(12) INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(19) World Intellectual Property Organization
International Bureau

(43) International Publication Date
26 August 2004 (26.08.2004)

PCT

(10) International Publication Number
WO 2004/073276 A1

(51) International Patent Classification[7]: H04L 27/26, 25/02, 1/06

(21) International Application Number: PCT/EP2003/001495

(22) International Filing Date: 14 February 2003 (14.02.2003)

(25) Filing Language: English

(26) Publication Language: English

(71) Applicant (*for all designated States except US*): **DOCOMO COMMUNICATIONS LABORATORIES EUROPE GMBH** [DE/DE]; Landsberger Strasse 308-312, 80687 München (DE).

(72) Inventor; and
(75) Inventor/Applicant (*for US only*): **AUER, Gunther** [DE/DE]; Westendstrasse 61, 80339 München (DE).

(74) Agent: **HOFFMANN . EITLE**; Patent- und Rechtsanwälte, Arabellastrasse 4, 81925 München (DE).

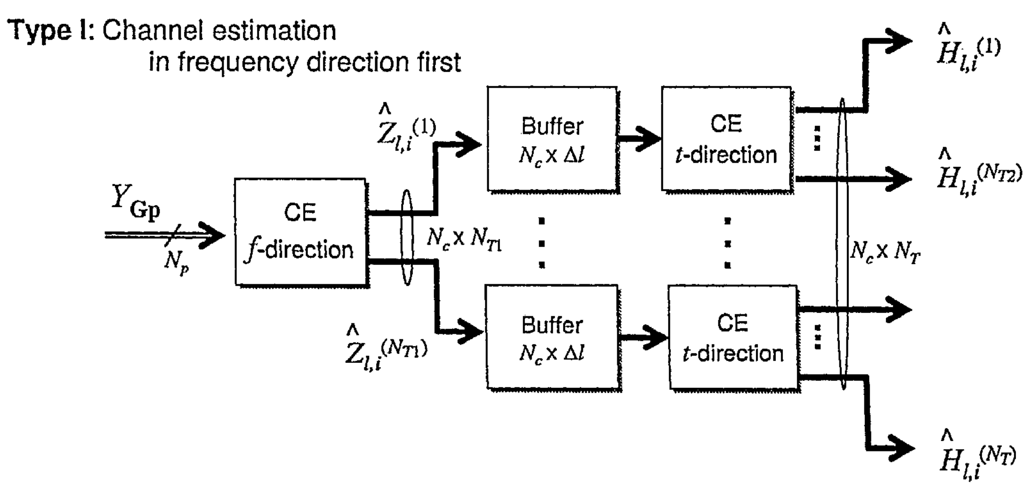
(81) Designated States (*national*): AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NO, NZ, OM, PH, PL, PT, RO, RU, SC, SD, SE, SG, SK, SL, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VC, VN, YU, ZA, ZM, ZW.

(84) Designated States (*regional*): ARIPO patent (GH, GM, KE, LS, MW, MZ, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, HU, IE, IT, LU, MC, NL, PT, SE, SI, SK, TR), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Published:
— *with international search report*

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

(54) Title: TWO-DIMENSIONAL CHANNEL ESTIMATION FOR MULTICARRIER MULTIPLE INPUT OUTPOUT COMMUNICATION SYSTEMS



(57) Abstract: The present invention extends the concept of two dimensional channel estimation to multicarrier multiple input multiple output communication systems using multicarrier modulated transmission signals impinging from a plurality of (NT) transmit antennas at the receiver side. It is assumed that the transmission signals carry a two dimensional data sequence with embedded pilot symbol. As basis of a two-stage channel estimation process the plurality of transmit antennas is divided into disjoint transmission antenna subsets (Amicrol); In a first stage of channel estimation impinging pilot sequences are separated in relation to transmission antenna subsets (Amicrol) by performing a first stage channel estimation to yield tentative estimates of a channel response in a first dimension of transmission. In a second stage of channel estimation impinging pilot sequences are separated in relation to antennas in transmission antenna subsets (Amicrol); by performing a second stage channel estimation for each antenna in each transmission antenna subset (Amicrol) to yield an estimation of the channel response.

WO 2004/073276 A1

TWO-DIMENSIONAL CHANNEL ESTIMATION FOR MULTICARRIER MULTIPLE INPUT OUTPUT COMMUNICATION SYSTEMS

Field of Invention

The present invention relates to channel estimation for multiple input multiple output communication systems, and in particular to a method and related channel estimator for multiple input multiple output communication systems using multi-carrier modulation schemes.

Background Art

The use of coherent transmission techniques in wireless communication systems requires the tracking of mobile radio channels, known as channel estimation. Since the signals transmitted from multiple transmit antennas are observed as mutual interference, channel estimation for multiple input multiple output MIMO communication systems is different from the single transmit antenna scenario. For multiple input multiple output MIMO communication systems using multi-carrier modulation schemes the received signal after multi-carrier demodulation is typically correlated in two dimensions, i.e. in time and frequency. In the following, orthogonal frequency division multiplexing OFDM will be referred to as one typical example for mutli-carrier moduilation schemes. The reason for this is that orthogonal frequency division multiplexing OFDM and variants thereof are the

most popular multi-carrier modulation schemes.

Communication systems employing multiple transmit and receive antennas, known as multiple input multiple output MIMO communication systems, can be used with orthogonal frequency division multiplexing OFDM to improve the communication capacity and quality of mobile radio systems. For orthogonal frequency division multiplexing OFDM communication systems with multiple transmit antennas, such as space-time codes as decibed in *A. Naguib, N.Seshadri, and A. Calderbank: "Space Time Coding and Signal Processing for High Data Rate Wireless Communications", IEEE Signal Processing Magazine, pp. 76-92, May 2000* or spacial multiplexing, different signals are transmitted form different transmit antennas simultaneously. Consequently, the received signal is the superposition of these signals, which implies challenges for channel estimation. Channel parameters are required for diversity combining, if space-time codes are used or alternatively for separation of superimposed signals if spatial multiplexing is used.

Approaches to such channel estimation are described in *Y. Li, N. Seshadri, and S. Ariyavisitakul: "Channel Estimation for OFDM Systems with Transmitte Diversity in Mobile Wireless Channels", IEEE Journal of Selected Areas on Communications, Vol 17., pp. 461-470, March 1999* and *Y. Li: "Simplified Channel estimation for OFDM Systems with Multiple Transmit Antennas", IEEE Transactions on Wireless Communications, Vol 1., pp. 67-75, January 2002*, in particular a channel estimation scheme for orthogonal frequency division multiplexing OFDM with multiple transmit antennas based on the difcrete fourier transform DFT.

Further, the estimators based on the least squares LS and minimum mean squared error MMSE criterion for OFDM-MIMO systems have been systematically derived in *Y. Gong and K. Letaief: "Low Rank Channel Estimation for Space-Time Coded Wideband OFDM Systems", Proc. IEEE Vehicular Technology Conference (VTC'2001-Fall), Atlantic City, USA, pp. 722-776, 2001.* Related solutions deal with one dimensional approaches where a known pilot OFDM symbol is followed by L data bearing OFDM symbols. This scheme is applicable for a quasi-static environment where the channel does not change significantly during L OFDM

symbols, i.e. indoor systems such as wireless local area networks WLAN.

To accomodate some mobility, the receiver may switch to decision directed channel estimation during the reception of the L data bearing OFDM symbols, as suggested in *Y. Li, N. Seshadri, and S. Ariyavisitakul: "Channel Estimation for OFDM Systems with Transmitte Diversity in Mobile Wireless Channels", IEEE Journal of Selected Areas on Communications, Vol 17., pp. 461-470, March 1999* and *Y. Li: "Simplified Channel estimation for OFDM Systems with Multiple Transmit Antennas"', IEEE Transactions on Wireless Communications, Vol 1., pp. 67-75, January 2002.* However, decision directed channel estimation which uses prior decisions of data symbols as pilot symbols, is significantly more complex than channel estimation schemes relying on pilots only.

Further, for OFDM-based systems with one transmit antenna, two dimensional channel estimation utilizing a scattered pilot grid can be employed, which satisfy the sampling theorem in time and frequency. For pilot-symbol aided channel estimation PACE known pilot symbols are multiplexed into the data stream. Interpolation is used to obtain the channel estimate for the information carrying symbols. PACE for single carrier systems was introduced in *J.K. Cavers: "An analysis of Pilot Symbol assited Modulation for Rayleigh Fading Channels", IEEE Transactions on Vehicular Technology, Vol. VT-40, pp. 686-693, November 1991.* In *R. Nilsson, O. Edfors, M. Sandell, and P. Boerjesson: "An Analysis of Two-Dimensional Pilot-Symbol Assisted Modulation for OFDM", Proc. IEEE Intern. Conf. on Personal Wireless Communications (ICPWC'97), Mumbai (Bombay), India, pp. 71-74, 1997* and *P. Hoeher, S. Kaiser, and P. Robertson: "Two-Dimensional Pilot-Symbol-Aided Channel Estimation by Wiener Filtering", Proc. IEEE Intern. Conf. on Acoustics, Speech, and Signal Processing.(ICASSP'97), Munich, Germany, pp. 1845-1848, 1997* two-dimensional 2D filtering algorithms have been proposed for pilot-symbol aided channel estimation PACE. However, such a 2D estimator structure is generally too complex for practical implementation.

To reduce the complexity, separating the use of time and frequency correlation has been proposed in *P. Hoeher, S. Kaiser, and P. Robertson: "Pilot-Symbol-Aided Channel Estimation in Time and Frequency", in Proc. Communication Theory*

Mini-Conference (CTMC) in con junction with IEEE Global Telecommunications Conference (GLOBECOM'97), Phoenix, USA, pp. 90-96, 1997. This combined scheme, termed double one-dimensional ($2 \times 1D$) pilot-symbol aided channel estimation PACE, uses separate Wiener filters, one in frequency direction and one in time direction.

Another approach to reduce the computational complexity is based on a transformation which concentrates the channel power to a few transform coefficients. Estimators based on the discrete Fourier transform DFT have the advantage that a computationally efficient transform in form of the FFT does exist, and that DFT-based interpolation is simple. In *Y. Li: "Pilot-Symbol-Aided Channel Estimation for OFDM in Wireless Systems", IEEE Transactions on Vehicular Technology, Vol. 49, pp. 1207-1215, July 2000,* the approach based on discrete Fourier transform DFT-based pilot-symbol aided channel estimation PACE was extended to two dimensional pilot-symbol aided channel estimation PACE by using the two dimensional FFT for the single antenna case.

However, a major problem for extending the approach to two dimensional channel estimation utilizing a scattered pilot grid to multiple input mulitple output MIMO communication systems is that the limitation of the number of transmit antennas which can be separated by a certain number of pilots. The minimum number of pilot symbols N_P' which are required to estimate N_T channel impulse responses (CIR) each of which having Q taps has been shown in *Y. Gong and K. Letaief: "Low Rank Channel Estimation for Space-Time Coded Wideband OFDM Systems", Proc. IEEE Vehicular Technology Conference (VTC'2001-Fall), Atlantic City, USA, pp. 722-776, 2001* to be

$$N_P' \geq N_T Q \qquad (1)$$

However, this means that the number of pilots required for channel estimation

grows with the number of transmit antennas N_T.

Summary of Invention

In view of the above, the object of the present invention is to extend the concept ot two dimensional channel estimation to MIMO systems.

According to the present invention this object is achieved through a method of two dimensional channel estimation for multiple input multiple output transmission systems using multicarrier modulated transmission signals impinging from a plurality of transmit antennas and carrying a two dimensional data sequence with embedded pilot symbols. In a first step the plurality of transmit antennas ís divided into disjoint transmission antenna subsets. In a second step impinging pilot sequences are seperated in relation to transmission antenna subsets by performing a first stage channel estimation to yield tentative estimates of a channel response in a first dimension of transmission. In a third step impinging pilot sequences are seperated in relation to antennas in transmission antenna subsets by performing a second stage channel estimation for each antenna in each transmission antenna subset to yield an estimation of the channel response.

An important advantage of the present invention is increased flexibilty in channel estimation. The reason for this is that by dividing the separation task of the superimposed transmission signals in time and frequency direction, a more efficient usage of the pilot symbols is possible. Hence, either the number of required pilot symbols may be reduced or the performance can be improved.

In view of the above, another important advantage of the present invention is the increase in the number of transmit antennas which can be estimated with a certain number of pilot symbols through application of a two stage channel estimaiton approach.

Yet another important advantage of the present invention is that the two stage channel estimation approach allows for tracking of channel variations even at high

Doppler frequencies. This is a prerequisite to support high velocities of mobile users and therefore to enable truely mobile multiple input multiple output MIMO communication systems.

According to a preferred embodiment of the present invention the first stage channel estimation is performed using pilot sequences arranged as a two dimensional grid of pilot symbols, wherein pilot symbols used for first stage channel estimation depend on the first dimension of transmission only and pilot symbols used for second stage channel estimation depend on a second dimension of transmission only. Preferably, pilot sequences are expressed in a product form for achieving seperability of pilot sequences in the first dimension of transmission and the second dimension of transmission.

An advantage of this preferred embodiment of the present invention is that utilization of a scattered pilot grid allows for efficient use of pilot symbols. Further, by matching the pilot spacing in time and frequency to the worst case channel characteristics higher mobile velocities can be supported with respect to conventional one dimensional schemes.

According to another preferred embodiment of the present invention, for a pilot spacing having a value of one the first stage channel estimation and/or the second stage channel estimation is achieved in a non-interpolating manner through yield of tentative estimates in relation to pilot symbol grid positions in the dimension of estimation. Alternatively, for a pilot spacing having a value larger than one the first stage channel estimation and/or the second stage channel estimation is achieved in an interpolating manner through yield of tentative estimates for all data sequence grid positions in the dimension of estimation.

An important advantage of this preferred embodiment is flexible support of different two dimensional pilot grids. In other words, the present invention may be flexibly applied using any type of pilot spacing, both, in frequncy and time

direction the application of suitable interpolation techniques.

Further preferred embodiments of the present invention relate selection of first dimension of transmission for the first channel estimation stage and the second channel estimation stage – i.e., in frequence direction or in time direction – and further to the selection of channel estimation domain – i.e., frequency domain channel estimation or time domain channel estimation. Here, according to the present invention any combination of dimension of transmission and channel estimation domain is supported.

The free selectability of dimension of transmission and channel estimation domain is further reason for flexibility of the channel estimation approach according to the present invention. It enables optimal consideration of multi-carrier related transmission parameters, selected pilot grid structure, and also application of computationanally most suitable channel estimation techniques.

According to another preferred embodiment of the present invention the channel estimation approach is applied to a celluar communication system with a frequency reuse factor of one such that base stations and related anntenna arrays form the plurality of transmit antennas and such that transmission antenna subsets and related transmission antennas are defined in relation to this plurality of transmit antennas.

An important advantage of this preferred embodiment of the present invention is the application of the two stage channel estimation techiques as outlined above to distributed antennas. In particular, it allows to handle a situation where a mobile user roams at a cell border. While data bearing symbols can be protected against interference using a channel code or spreading, this is not possible for pilot symbols. According to the present invention through appropriate definition of subset in relation to cells in the celluar communication system.

According to yet another preferred embodiment of the present invention there is provided a computer program product directly loadable into the internal memory

of a channel estimator for estimating multiple input multiple output transmission channels in two dimensions comprising software code portions for performing the steps of the method of two dimensional channel estimation according to the present invention when the product is run on a processor of the channel estimator.

Therefore, the present invention is also provided to achieve an implementation of the inventive method steps on computer or processor systems. In conclusion, such implementation leads to the provision of computer program products for use with a computer system or more specifically a processor comprised, e.g., in a channel estimator for estimating multiple input multiple output transmission channels in two dimensions.

The programs defining the function of the present invention can be delivered to a computer/processor in many forms, including, but not limited to information permanently stored on non-writeable storage media, e.g., read only memory devices such as ROM or CD ROM discs readable by processors or computer I/O attachments; information stored on writable storage media, i.e. floppy discs and hard drives; or information convey to a computer/processor through communication media such as local area network and/or telephone networks and/or Internet or other interface devices. It should be understood, that such media when carrying processor readable instructions implementing the inventive concept represent alternate embodiments of the present invention.

Description of Drawing

In the following the best mode and preferred embodiments of the present invention will be explained with reference to the drawing in which:

Fig. 1 shows a schematic diagram of an OFDM based multiple input multiple output MIMO communication system for explanation of the system model under-

lying the present invention;

Fig. 2 shows a schematic diagram illustrating OFDM modulation and demodulation, respectively;

Fig. 3 shows a scattered pilot grid suitable for two dimensional channel estimation according to the present invention;

Fig. 4 shows a schematic diagram of a channel estimator for estimating multiple input multiple output transmission channels of mutlicarrier communication systems according to the present invention;

Fig. 5 shows a flowchart of operation of the channel estimator shown in Fig. 5;

Fig. 6 shows a schematic diagram illustrating the principle of 2x1D channel estimation underlying the present invention;

Fig. 7 shows a schematic diagram of a channel estimator for estimating multiple input multiple output transmission channels of mutlicarrier communication systems according to the present invention, wherein channel estimation is performed in frequency direction first;

Fig. 8 shows a further scattered pilot grid suitable for two dimensional channel estimation according to the present invention;

Fig. 9 shows a further scattered pilot grid corresponding to an digital video broadcast DVB-T application and suitable for two dimensional channel estimation according to the present invention;

Fig. 10 shows a schematic diagram of a channel estimator for estimating multiple input multiple output transmission channels of mutlicarrier communication systems according to the present invention, wherein channel estimation is per-

formed in time direction first;

Fig. 11 shows a schematic diagram of an estimator stage adapted to achieve channel estimation in the time domain according to the present invention;

Fig. 12 shows a schematic diagram of a further estimator stage adapted to achieve channel estimation in the time domain according to the present invention; and

Fig. 13 shows an application of the two stage channel estimation approach according to the present invention to a cellular communication system with a frequency reuse factor of one.

Description of Best Mode and Preferred Embodiments

In the following, the best mode and preferred embodiments of the present invention will be explained with reference to the drawing. Initially, some basic considerations underlying differential multiple-length transmit diversity and related diversity reception will be explained for a better understanding of the present invention.

Basic Considerations

The present invention addresses pilot-symbol aided channel estimation PACE schemes for multi-carrier multiple input multiple output MIMO communication systems which are based on the insertion of a pilot grid. Typically for a multi-carrier multiple input multiple output MIMO communication system, pilot-symbol aided channel estimation PACE – also referred to as PACE in the following – is applied across subcarriers in frequency direction and over several multi-carrier transmission symbols – e.g., OFDM symbols - in two dimensions, resulting in two

dimensional 2D PACE.

The present invention as explained in the following is not restricted to a particual type of multi-carrier multiple input multiple output MIMO communication system, and may be applied, e.g., to orthogonal frequency divsion multiplexing OFDM, discrete multitone transmission DMT, filtered multitone transmission FMT, or biorthogonal frequency division multiplexing BFDM.

Another multi-carrier multiple input multiple output MIMO communication system is multi-carrier code divsion mutliple access MC-CDMA where spreading in frequency and/or time direction is introduced in addition to the multi-carrier modulation. Yet another multi-carrier multiple input multiple output MIMO communication system is a multi-carrier code divsion mutliple access MC-CDMA system with a variable spreading factor, namely variable spreading factor orthogonal frequency and code division multiple access VSF–OFCDM.

From the above, it should be understood that pilot symbol aided channel estimation PACE may be applied to all multi-carrier multiple input multiple output MIMO communication systems operating with transmission signals being correlated in two dimensions. Therefore, all these multi-carrier multiple input multiple output MIMO communication systems may employ the different embodiments of the present invention as explained in the following.

System Model

Fig. 1 shows a schematic diagram of an OFDM based multiple input multiple output MIMO communication system – also referred to as OFDM system in the following - for explanation of the system model underlying the present invention. Further, Fig. 2 shows a schematic diagram illustrating OFDM modulation and demodulation in correspondence to the OFDM based multiple input multiple output MIMO communication system shown in Fig. 1.

As shown in Fig. 1, for the considered OFDM-based MIMO system, one

OFDM modulator is employed on each transmit antenna, while OFDM demodulation is performed independently for each receive antenna. The signal stream is divided into N_c parallel substreams, typically for any multi-carrier modulation scheme. The i^{th} substream, also referred to as subcarrier in the following, of the ℓ^{th} symbol block, named OFDM symbol in the following, is denoted by $X_{\ell,i}$. An inverse DFT with N_{FFT} points is performed on each block, and subsequently the guard interval having N_{GI} samples is inserted to obtain $x_{\ell,n}$. After D/A conversion, the signal $x(t)$ is transmitted over a mobile radio channel with response $h(t,\tau)$.

As shown in Fig. 1, the received signal at receive antenna ν consists of superimposed signals from N_T transmit antennas. Assuming perfect synchronization, the received signal of the equivalent baseband system impinging at receive antenna ν at sampling instants $t = [n + \ell N_{sym}]T_{spl}$ is in the form

$$y^{(\nu)}_{\ell,n} \triangleq y^{(\nu)}([n+\ell N_{sym}]T_{spl}) = \sum_{\mu=1}^{N_T} \int_{-\infty}^{\infty} h^{(\mu,\nu)}(t,\tau) \cdot x^{(\mu)}(t-\tau)\, d\tau + n(t) \Bigg|_{t=[n+\ell N_{sym}]T_{spl}}$$

where $x^{(\mu)}(t)$ denotes transmitted signal of transmit antenna μ after OFDM modulation, $n(t)$ represents additive white Gaussion noise, and $N_{sym} = N_{FFT} + N_{GI}$ accounts for the number of samples per OFDM symbol.

As shown in Fig. 2, at the receiver the guard interval is removed and the information is recovered by performing an DFT on the received block of signal samples, to obtain the output of the OFDM demodulation $Y_{\ell,i}$. The received signal at receive antenna ν after OFDM demodulation given by

$$Y^{(\nu)}_{\ell,i} = \sum_{\mu=1}^{N_T} X^{(\mu)}{}_{\ell,i}\, H^{(\mu,\nu)}_{\ell,i} + N_{\ell,i} \qquad (2)$$

where $X^{(\mu)}{}_{\ell,i}$ and $H^{(\mu,\nu)}_{\ell,i}$ denotes the transmitted information symbol and the channel transfer function (CTF) of transmit antenna μ, at subcarrier i of the ℓ^{th} OFDM symbol, respectively. The term $N_{\ell,i}$ accounts for additive white Gaussian noise AWGN with zero mean and variance N_0. It is assumed that the transmitted signal consists of L OFDM symbols, each having N_c subcarriers.

Channel Characteristics

While in the following, a way to model channel characteristics will be explained, it is important to note that the present invention is not restricted in any way through such a model. To the contrary the present invention is applicable to actually existing mobile radio channels. For the application of the present invention it is only of relevance that mobile radio channels are band limited. Preferably, they should also be be limited in time, which practically is the case in view of maximum delay of mobile radio channel. Further, preferably different transmit antennas and receive antennas should be uncorrelated.

The present invention considers a time-variant frequency selective fading channel. The number of non-zero taps is typically smaller or equal to the maximum delay of the channel, $Q_0 \leq Q$. The channel impulse response CIR between transmit antenna μ and receive antenna ν is defined by

$$h^{(\mu,\nu)}(t,\tau) = \sum_{q=1}^{Q_0} h_q^{(\mu,\nu)}(t) \cdot \delta(\tau - \tau_q^{(\mu,\nu)}) \tag{3}$$

where $h_q^{(\mu,\nu)}(t)$ and $\tau_q^{(\mu,\nu)}$ are the complex amplitude and delay of the q^{th} channel tap.

According to the present invention it is assumed that the Q_0 channel taps and all antennas are mutually uncorrelated. The channel taps $h_q^{(\mu,\nu)}(t)$ are zero-mean complex independent identically distributed (i.i.d.) Gaussian random variables. Due to the motion of the vehicle $h_q^{(\mu,\nu)}(t)$ will be time-variant caused by the Doppler effect. The q^{th} channel tap $h_q^{(\mu,\nu)}(t)$ is a wide sense stationary WSS Gaussian process, being band-limited by the maximum Doppler frequency ν_{max}.

Further, it is commonly assumed that the channel impulse response CIR is approximately constant during one OFDM symbol, so the time dependency of the CIR within one OFDM symbol can be dropped, i.e. $h_q^{(\mu,\nu)}(t) \approx h_{q,\ell}^{(\mu,\nu)}$ for $t \in [\ell T, (\ell+1)T]$. Although this is strictly true only for time-invariant channels,

this assumption seems to be most often justified in practice, and a good system design should ensure that the OFDM symbol duration is sufficiently short.

Further, the channel transfer function CTF of equation (2), is the the Fourier transform of the CIR $h^{(\mu,\nu)}(t,\tau)$. Sampling the result at time $t = \ell T_{sym}$ and frequency $f = i/T$, the CTF at subcarrier i of OFDM symbol ℓ becomes

$$H_{\ell,i}^{(\mu,\nu)} = H^{(\mu,\nu)}(\ell T_{sym}, i/T) = \sum_{q=1}^{Q_0} h_{q,\ell}^{(\mu,\nu)} \, e^{-j2\pi \tau_q i/T} \tag{4}$$

where $T_{sym} = (N_{FFT} + N_{GI})T_{spl}$ and $T = N_{FFT}T_{spl}$ represents the OFDM symbol duration with and without the guard interval, respectively. The matrix form of equation (4) is given by

$$\mathbf{H}_{\ell}^{\prime(\mu,\nu)} = \mathbf{T}^{(\mu,\nu)}\mathbf{h}_{\ell}^{\prime(\mu,\nu)} \tag{5}$$

where $\mathbf{T}^{(\mu,\nu)}$ represents the transformation matrix which transforms $\mathbf{h}_{\ell}^{\prime(\mu,\nu)}$ into the frequency domain, defined by

$$\left\{\mathbf{T}^{(\mu,\nu)}\right\}_{q,i} = \exp\left(-j2\pi \frac{\tau_q^{(\mu,\nu)} i}{T_{spl}N_{FFT}}\right) \; ; \quad 0 \le i \le N_c - 1 \, , \quad 1 \le q \le Q_0 \tag{6}$$

of dimension $N_c \times Q_0$.

Further, if the guard interval is longer than the maximum delay of the channel, i.e. $N_{GI} \geq Q$, where $Q \geq Q_0$ denotes the total number of channel taps, the orthogonality at the receiver after OFDM demodulation is maintained, and the received signal of equation (2) is obtained.

Assuming the fading at the receiver antennas is mutually uncorrelated, the channel estimation according to the present invention will be performed independently for each antenna. Therefore, in the description to follow the marker for the receive antenna ν is omitted.

Further, according to the present invention a channel is defined to be sample spaced if the tap delays τ_q are multiples of the sample instant T_{spl}, i.e.

$$\tau_q = T_{spl}\,\beta_q \; ; \qquad 1 \le q \le Q_0 \, , \quad \beta_q \in \{0, 1, \cdots, Q\} \tag{7}$$

where β_q is an arbitrary integer, equal to or larger than zero. In this case $H_{\ell,i}^{(\mu,\nu)}$ in equation (4) becomes the DFT of the CIR$h_{\ell,n}^{(\mu,\nu)}$.

In view of the above, $H_{\ell,i}^{(\mu,\nu)}$ can be expressed in matrix notation

$$\mathbf{H}_{\ell}'^{(\mu,\nu)} = \mathbf{F}\mathbf{h}_{\ell}'^{(\mu,\nu)} \tag{8}$$

where $\mathbf{F}$ represents the DFT-matrix of dimension $N_{GI} \times N_c$, defined by

$$\{\mathrm{F}\}_{n,i} = e^{-j2\pi ni/N_{\mathrm{FFT}}}; \qquad 0 \leq i \leq N_c - 1, \quad 0 \leq n \leq N_{GI} - 1 \tag{9}$$

While for a real channel the tap delays τ_q will not be multiples of the sample duration T_{spl}, however, for many applications the sample spaced channel model does approximate a real channel sufficiently well.

Principle of Pilot Symbol Aided Channel Estimation for OFDM-based MIMO Systems

Fig. 3 shows a scattered pilot grid suitable for two dimensional channel estimation according to the present invention.

As shown in Fig. 3, pilot aided channel estimation PACE is based on periodically inserting known symbols, termed pilot symbols in the data sequence. If the spacing of the pilot symbols is sufficiently close to satisfy the sampling theorem, channel estimation and interpolation for the entire data sequence is possible.

For two dimensional pilot aided channel estimation in the sense of the present invention it must be taken into account that for OFDM the fading fluctuations are in two dimensions, in time and frequency. In order to satisfy the two-dimensional sampling theorem, the pilot symbols are therefore scattered throuout the time-frequency grid, which yields a two-dimensional pilot grid. Another reason for selecting scattered pilot grids is to maximize spectral efficiency.

Description of Pilot Grids in Two Dimensions

To formally describe a regular grid in the 2D plane according to the present invention the total number of pilots transmitted in one frame is defined to $N_P = N_P' \cdot N_P''$, with $N_P' = N_c/D_f$ and $N_P'' = L/D_t$ being the number of pilots in frequency and time direction respectively. Here, the following notation is used: given a matrix describing a 2D structure $\mathbf{X}$, the subsets which describe the dimension corresponding to the frequency and time directions are denoted by $\mathbf{X}'$ and $\mathbf{X}''$, respectively.

Denoting the pilot symbol of subcarrier i and OFDM symbol ℓ by $\mathbf{p} = [i, \ell]^T$, any regular 2D grid for use in the framework of the present invention is be described by

$$\left\{\mathcal{G} : \quad \mathbf{p} = \mathbf{G}\tilde{\mathbf{p}} + \mathbf{g}_0, \quad \forall\, \tilde{\mathbf{p}} \in \mathbb{I}^{N_P' \times N_P''}\right\} \tag{10}$$

where $\tilde{\mathbf{p}} = [\tilde{i}, \tilde{\ell}]^T$ denotes the index of the $\tilde{i}^{\text{th}}$ and $\tilde{\ell}^{\text{th}}$ pilot in frequency and time direction, respectively, and $\mathbf{g}_0$ defines a pilot grid offset.

For multiple input multiple output MIMO communication systems – also referred to as MIMO system in the follwoing – every transmission signal at transmit antenna may to use its own pilot grid. This enables the receiver to separate the superimposed transmission signals from different transmission antennas.

To describe pilot symbol-assisted channel estimation in the sense of the present invention there is defined a subset of the received signal sequence containing only the pilots, $\{\tilde{X}^{(\mu)}_{\tilde{\ell},\tilde{i}}\} = \{X^{(\mu)}_{\mathbf{G}\tilde{\mathbf{p}}}\}$, with $\mathbf{G}\tilde{\mathbf{p}} = [i, \ell]^T \in \mathcal{G}$, sampled at a D_f times lower rate $\tilde{i} = \lfloor i/D_f \rfloor$ in frequency direction, and at a D_t times lower rate $\tilde{\ell} = \lfloor \ell/D_t \rfloor$ in time direction, respectively. As a general convention, variables describing pilot symbols will be marked with a $\tilde{}$ in the following description.

Further, without restricting scope of protection, it may be assumed that the

pilots $\tilde{X}^{(\mu)}{}_{\tilde{\ell},\tilde{i}}$ are chosen from a PSK constellation, so $|\tilde{X}^{(\mu)}_{\tilde{\ell},\tilde{i}}| = 1$.

For the particular example shown in Fig. 3 the structure of the pilot grid is defined by G which is

$$\mathrm{G} = \begin{bmatrix} D_f & 0 \\ 0 & D_t \end{bmatrix}$$

where D_f and D_t denotes the so-called pilot spacing in frequency and time, respectively. For the pilot grid shown in Fig. 3 the particular values are $D_f = 5$ and $D_t = 5$.

It should be noted that before transmission, the pilots $\{\tilde{X}^{(\mu)}_{\tilde{\ell},\tilde{i}}\}$ may be multiplied by an outer pilot sequence $\{\tilde{X}_{0_{\tilde{\ell},\tilde{i}}}\}$ which is identical for all transmit antennas to yield the transmitted pilot sequence

$$\tilde{X}^{(\mu)}_{T_{\tilde{\ell},\tilde{i}}} = \tilde{X}_{0_{\tilde{\ell},\tilde{i}}} \cdot \tilde{X}^{(\mu)}_{\tilde{\ell},\tilde{i}}$$

Further, the outer pilot sequence $\{\tilde{X}_{0_{\tilde{\ell},\tilde{i}}}\}$ is chosen to have a low peak to power average ratio in the time domain and/or to have good correlation properties for synchronization.

As shown in Fig. 2, at the receiver the cyclic prefix is removed and an FFT is performed to yield the received signal after OFDM demodulation. Assuming perfect synchronization, the received signal $Y_{\ell,i}$ of equation (2) is obtained. For channel estimation the received signal at the pilot positions are demultiplexed from the data stream, and after removing the outer pilot sequence, by dividing through $\tilde{X}_{0_{\tilde{\ell},\tilde{i}}}$, the received pilot is obtained according to

$$\tilde{Y}_{\tilde{\ell},\tilde{i}} \triangleq Y_{\mathrm{G}\tilde{\mathrm{p}}} = \sum_{\mu=1}^{N_T} X^{(\mu)}_{\mathrm{G}\tilde{\mathrm{p}}} H^{(\mu)}_{\mathrm{G}\tilde{\mathrm{p}}} + N_{\mathrm{G}\tilde{\mathrm{p}}} \quad (11)$$

where $\mathrm{G}\tilde{\mathrm{p}}$ is defined in (10).

Basic Consideration for 2x1D-Pilot Assisted Channel Estimation for Multiple Input Multiple Ouput Communication Systems

Fig. 4 shows a schematic diagram of a channel estimator for estimating multiple input multiple output transmission channels of mutlicarrier communication systems according to the present invention.

As shown in Fig. 4, the channel estimatior 10 comprises a first estimator stage 12 and a second estimator stage 14. Further, the channel estimator comprises a transmission antenna subset memory 16.

Fig. 5 shows a flowchart of operation of the channel estimator shown in Fig. 4.

As shown in Fig. 5, operation of the channel estimator according to the presen invention relies on a method where in a firsat step S10 the plurality of transmit antennas into disjoint transmission antenna subsets. Then, in a step S12 impinging pilot sequences in relation to transmission antenna subsets are seperated by performing a first stage channel estimation to yield tentative estimates of a channel response in a first dimension of transmission. Then, in a step S14 impinging pilot sequences in relation to antennas in transmission antenna subsets by performing a second stage channel estimation for each antenna in each transmission antenna subset to yield an estimation of the channel response.

Fig. 6 shows a schematic diagram illustrating the principle of channel estimationg according to the present invention.

As shown in Fig. 6 and outlined above, in the step S12 channel estimation is performed in one dimension, yielding tentative estimates for all subcarriers of these OFDM symbols. These tentative estimates are then used in step S14 as new pilots, in order to estimate the channel for the entire frame.

As shown in Fig. 6, the second stage does not only interpolate between OFDM symbols having pilots, but it does also improve the accuracy of the tentative esti-

mates.

Further, according to the present invention Either channel estimation in frequency direction or time direction may be performed first. Reference to the case where channel estimation in frequency direction is performed first will be made by $2\times$1D-PACE type I in the following. Further, reference to the case where channel estimation in time direction is performed first will be mde by $2\times$1D-PACE type II in the following.

To formally describe the problem, the received pilot of OFDM symbol ℓD_t is considered at the $(iD_f)^{\text{th}}$ subcarrier

$$Y_{\tilde{\ell}D_t,\tilde{i}D_f} = \sum_{\mu=1}^{N_T} X^{(\mu)}_{\tilde{\ell}D_t,\tilde{i}D_f} H^{(\mu)}_{\tilde{\ell}D_t,\tilde{i}D_f} + N_{\tilde{\ell}D_t,\tilde{i}D_f} \qquad \tilde{\ell} = \{1,2,\cdots,L/D_t\}$$

$$\tilde{i} = \{1,2,\cdots,N_c/D_f\}$$

where $X^{(\mu)}{}_{\tilde{\ell}D_t,\tilde{i}D_f}$ and $H^{(\mu,\nu)}_{\tilde{\ell}D_t,\tilde{i}D_f}$ denotes the transmitted pilot symbol and the channel transfer function (CTF) of transmit antenna μ, at subcarrier $i = \tilde{i}D_f$ of the $\ell = \tilde{\ell}D_t{}^{\text{th}}$ OFDM symbol, respectively.

Further, it is assumed that the CTF varies in the ℓ and in the i variable, i.e. in time and in frequency. The term $N_{\ell D_t,iD_f}$ accounts for additive white Gaussian noise AWGN. L represents the number of OFDM symbols per frame, N_c is the number of subcarriers per OFDM symbol, D_f and D_t denote the pilot spacing in frequency and time, and N_T is the number of transmit antennas.

The overall object underlying the present invention is to estimate $H^{(\mu)}_{\ell,i}$ for all $\{\ell, i, \mu\}$ within the frame.

To achieve this object, generally for MIMO channel estimation, in addition to channel estimation and interpolation it is also necessary to separate the impinging signals from N_T transmit antennas. Dividing the signals corresponding to the N_T transmit antennas into N_{T1} subsets, yields N_{T1} groups each of which having N_{T2} signals, such that $N_T = N_{T1}N_{T2}$. In other words, for a MIMO system having N_T

transmit antennas, according to the present invention signals corresponding to N_{T2} transmit antennas into one set, to form N_{T1} subsets.

In other words, the basic concept underlying the present invention is to divide the separation task into two stages, in the way that in step S12 channel estimation is performed in one dimension, at OFDM symbols $\ell := \tilde{\ell} D_t$, yielding tentative estimates for all subcarriers of these OFDM symbols. The second step S14 uses these tentative estimates as new pilots, in order to estimate the channel for the entire frame.

As shown in Fig. 6, the second stage estimation does not only interpolate between OFDM symbols having pilots, but it does also improve the accuracy of the tentative estimates. Therefore, the different embodiments of the present invention as explained in detail in the following have significantly reduced complexity while there is little degradation in performance.

According to Fig. 6 channel estimation in frequency direction is performed first, to reverse the order such that channel estimation in time direction is performed first is straightforward. In the following reference will be made to the case where channel estimation in frequency direction is performed first by 2×1D-PACE type I. to the contrary, reference to the case where channel estimation in time direction is performed first will be made to as 2×1D-PACE type II.

Considering channel estimation in time direction, it is proposed to use the pilots $[Y_{1,i}, Y_{D_t,i}, \cdots, Y_{D_t N''_P+1,i}] \in \mathcal{G}$, in order to estimate $\widehat{H}^{(\mu)}_{\ell,i}$. Further, for smoothing type filtering it is proposed to use past as well as future pilots to estimate $\widehat{H}^{(\mu)}_{\ell,i}$, this means $1 < \ell < D_t N''_P$. Therefore, for smoothing an estimate cannot be obtained until all pilots have been received, which requires buffering of $\Delta\ell = D_t N''_P - \ell$ OFDM symbols.

As will be shown in more detail in the follwoing, the alternative is to use prediction type filtering where $\ell > D_t N''_P$. In this case only past pilots are used for

channel estimation in time direction.

Obviously, prediction type filtering does not require any buffering, however, the performance with respect to smoothing degrades. For channel estimation in frequency direction, on the other hand, all pilots of one OFDM symbol are being received together, so no buffering is required. However, the accuracy of the channel estimates typically degrades near the band edges.

2×1D-PACE type I

Fig. 7 shows a schematic diagram of a channel estimator for estimating multiple input multiple output transmission channels of mutlicarrier communication systems according to the present invention, wherein channel estimation is performed in frequency direction first.

According to the present invention it is prposed to extend 2×1D-PACE to OFDM-based MIMO channel estimation. For MIMO channel estimation it is necessary to separate the impinging signals from N_T transmit antennas. Let the MIMO system having N_T transmit antennas be denoted by set the $\mathcal{A}$. Further, according to the present invention the set of N_T transmit antennas is devided into N_{T1} subsets $\mathcal{A}_{\mu_1} \subset \mathcal{A}$, with $\mu_1 = \{1, \cdots, N_{T1}\}$. Each subset contains N_{T2} transmit antennas, such that $N_T = N_{T1}N_{T2}$.

In other words, for a MIMO system having N_T transmit antennas, according to the present invention it is proposed to group the signals corresponding to N_{T2} transmit antennas into one set, to form N_{T1} subsets of $\mathcal{A}$. Without restricting scope of protection, one may assume that all sets have the same number of transmit antennas, and the subsets $\mathcal{A}_{\mu_1}$ are disjoint, i.e. each transmit antenna can only be in one set.

The concept underlying the present invention is to divide the separation task into two stages, in the way that we first separate a subset of the $N_{T1} < N_T$ signals together with channel estimation in the first estimation stage. The resulting signal

$\widehat{Z}_{\tilde{\ell},i}^{(\mu_1)}$ will be a superposition of $N_{T2} = N_T/N_{T1}$ signals.

In the second estimation stage the remaining N_{T2} superimposed signals are seperated for each of the N_{T1} signals of the first estimation stage, together with channel estimation in the second dimension, to yield the estimate of the frequency response $\widehat{H}_{\tilde{\ell},i}^{(\mu)}$.

Fig. 7 illustrates the basic idea for type I of the proposed scheme.

It should be noted, that the buffer shown in Fig. 7 is used in order to apply smoothing type filtering in time direction. However and as outlined above, the present invention is applicable to, both, smoothing and prediction type filtering, so the buffer shown in Fig. 7 is optional and depends on the particular channel estimation algorithm being used.

Fig. 8 and 9 show pilot sequence designs which may be used to support the two stage approach for OFDM-based MIMO channel estimation accordance to the present invention. Here, the pilot grid shown in Fig. 9 corresponds to the DVB-T pilot grid according to ETSI EN 300 744, V 1.4.1 (2001-01). Further standards – however, not to be considered as restricting scope of protection – would be IEEE 802.11a or ETSI TS 101 475 HIPERLAN/2.

In a more genral sense, in order to extend the pilot sequence design for an OFDM-based system with multiple transmit antennas, according to a preferred embodiment of the present invention the pilot sequence of transmit antenna μ is defined by $\{\tilde{X}_{\tilde{\ell},\tilde{i}}^{(\mu)}\}$.

It is proposed to choose a pilot sequences that can be expressed in the product form

$$\tilde{X}_{\tilde{\ell},\tilde{i}}^{(\mu)} = \tilde{X}_{1_{\tilde{i}}}^{(\mu_1)} \cdot \tilde{X}_{2_{\tilde{\ell}}}^{(\mu_2)}, \quad \mu = \mu_2 + N_{T2} \cdot (\mu_1 - 1), \quad \begin{array}{l} \mu_1 = \{1, \cdots, N_{T1}\} \\ \mu_2 = \{1, \cdots, N_{T2}\} \end{array}$$
$$1 \le \tilde{i} \le N_P', \quad 1 \le \tilde{\ell} \le N_P'' \tag{12}$$

where $\tilde{X}_{1_{\tilde{i}}}^{(\mu_1)}$ and $\tilde{X}_{2_{\tilde{\ell}}}^{(\mu_2)}$ are the pilot symbols for the first and second stage, respectively.

It should be noted that the pilot symbol of the first stage $\tilde{X}_{1_{\tilde{i}}}^{(\mu_1)}$ only depends on the subcarrier index $\tilde{i}$, while the pilot symbol of the second stage $\tilde{X}_{2_{\tilde{\ell}}}^{(\mu_2)}$ only depends on OFDM symbol $\tilde{\ell}$.

This notation implies that $\{\tilde{X}_{1_{\tilde{i}}}^{(\mu_1)}\}$ is identical for all OFDM symbols, independent of ℓ. Correspondingly, the sequence $\{\tilde{X}_{2_{\tilde{\ell}}}^{(\mu_2)}\}$ which accounts for the pilots of one subcarrier, is also independent of the subcarrier index i.

Preferably, the pilot sequences $\{\tilde{X}_{1_{\tilde{i}}}^{(\mu_1)}\}$ and $\{\tilde{X}_{2_{\tilde{\ell}}}^{(\mu_2)}\}$ are chosen from orthogonal designs, e.g., Walsh sequences or phase shifted sequences.

Substituting the proposed 2D pilot sequence of (12) into the received pilots in (11), the following is obtained

$$\begin{aligned} \tilde{Y}_{\tilde{\ell},\tilde{i}} &= \sum_{\mu_1=1}^{N_{T1}} \sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{1_{\tilde{i}}}^{(\mu_1)} \tilde{X}_{2_{\tilde{\ell}}}^{(\mu_2)} \tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu_2 + N_{T2}\cdot(\mu_1-1))} + \tilde{N}_{\tilde{\ell},\tilde{i}} \\ &= \sum_{\mu_1=1}^{N_{T1}} \tilde{X}_{1_{\tilde{i}}}^{(\mu_1)} \underbrace{\sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_{\tilde{\ell}}}^{(\mu_2)} \tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu_2 + N_{T2}\cdot(\mu_1-1))}}_{\tilde{Z}_{\tilde{\ell},\tilde{i}}^{(\mu_1)}} + \tilde{N}_{\tilde{\ell},\tilde{i}} \end{aligned} \tag{13}$$

where $\tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu)}$ is the frequency response of transmit antenna μ.

Further, the received pilot sequence of subset $\mathcal{A}_{\mu_1}$ is given by

$$\tilde{Z}_{\tilde{\ell},\tilde{i}}^{(\mu_1)} \triangleq \sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_{\tilde{\ell}}}^{(\mu_2)} \tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu_2 + N_{T2}\cdot(\mu_1-1))} \tag{14}$$

In view of the above, the task of the first estimation stage is to estimate $\widehat{Z}_{\tilde{\ell},\tilde{i}}^{(\mu_1)}$; that is to separate the N_{T1} groups, and then to estimate and interpolate the channel

in frequency direction.

It should be noted that the pilot sequence $\tilde{X}^{(\mu_2)}_{2_{\tilde{\ell}}}$ is constant with respect to the subcarrier index $\tilde{i}$. This means that $\tilde{Z}^{(\mu_1)}_{\tilde{\ell},\tilde{i}}$ is a superposition of N_{T2} waveforms $\tilde{H}^{(\mu)}_{\tilde{\ell},\tilde{i}}$ multiplied with a constant phase term $\tilde{X}^{(\mu_2)}_{2_{\tilde{\ell}}}$.

Further, the outputs of the first estimation stage are subsequently used as inputs for the second estimation stage shown 7.

As shown in Fig. 7, in the second estimation stage the channel is estimated in time direction to separate the remaining N_{T2} signals per subset to yield the estimate of the frequency response $\widehat{H}^{(\mu)}_{\ell,i}$.

Using this framework, according to the present invention it is üroposed to apply available one dimensional schemes for OFDM-based MIMO systems to perform channel estimation.

2×1D-PACE type II

Fig. 10 shows a schematic diagram of a channel estimator for estimating multiple input multiple output transmission channels of multicarrier communication systems according to the present invention, wherein channel estimation is performed in time direction first.

As shown in Fig. 10, the major difference of 2×1D-PACE type II over 2×1D-PACE type I is that that the separation N_{T1} subsets in the first estimation stage is performed in conjunction with channel estimation in time direction. This yields for the pilot symbols of 2×1D-PACE type II:

$$\tilde{X}^{(\mu)}_{\tilde{\ell},\tilde{i}} = \tilde{X}^{(\mu_1)}_{1_{\tilde{\ell}}} \cdot \tilde{X}^{(\mu_2)}_{2_{\tilde{i}}}\,, \quad \mu = \mu_2 + N_{T2}\cdot(\mu_1 - 1)\,, \quad \begin{array}{l} \mu_1 = \{1, \cdots, N_{T1}\} \\ \mu_2 = \{1, \cdots, N_{T2}\} \end{array}$$
$$1 \le \tilde{i} \le N'_P\,, \quad 1 \le \tilde{\ell} \le N''_P \tag{15}$$

From the above, it may be seen that the pilot sequences are very similar to 2×1D-PACE type I in equation (12), the only difference is that the subcarrier index $\tilde{i}$ and the OFDM symbol index $\tilde{\ell}$ are exchanged. Substituting the proposed 2D pilot sequence of equation (15) into the received pilots in equation (11), the received pilot sequence can be represented according to

$$\tilde{Y}_{\tilde{\ell},\tilde{i}} = \sum_{\mu_1=1}^{N_{T1}} \tilde{X}_{1_{\tilde{\ell}}}^{(\mu_1)} \underbrace{\sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_{\tilde{i}}}^{(\mu_2)} \tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu_2 + N_{T2} \cdot (\mu_1 - 1))}}_{\tilde{Z}_{\tilde{\ell},\tilde{i}}^{(\mu_1)}} + \tilde{N}_{\tilde{\ell},\tilde{i}} \tag{16}$$

where again the subcarrier index $\tilde{i}$ and the OFDM symbol index $\tilde{\ell}$ are exchanged with respect to (13).

This allows the separation of the N_{T1} subsets in time direction. The received pilot sequence of subset $\mathcal{A}_{\mu_1}$ is defined by

$$\tilde{Z}_{\tilde{\ell},\tilde{i}}^{(\mu_1)} \triangleq \sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_{\tilde{i}}}^{(\mu_2)} \tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu_2 + N_{T2} \cdot (\mu_1 - 1))} \tag{17}$$

In view of he above and as shown in Fig. 10, the task of the first estiamtion stage is to estimate $\widehat{Z}_{\ell,\tilde{i}}^{(\mu_1)}$; that is to separate the N_{T1} groups, and then to estimate and interpolate the channel in time direction, i.e. over the ℓ variable.

As shown in Fig. 10, if smoothing type filtering is used, $\Delta\ell D_t$ OFDM symbols need to be buffered in order to estimate $\widehat{Z}_{\ell,\tilde{i}}^{(\mu_1)}$. For the second estimation stage it is proposed estimate the channel in frequency direction to separate the remaining N_{T2} signals per subset.

Comparison between 2×1D-PACE type I and type II

It can be shown that the performance of both schemes type I and type II is identical for most implementations. However, the computational complexity, in terms of number of multiplications required for channel estimation may differ. The actual computational complexity very much depends on the system parameters, the pilot grid structure and the channel estimation algorithm which is used. Furthermore, the DSP or hardware architecture may favor one scheme.

More importantly however, does the selection of a certain pilot grid rule out the implementation of a particular scheme. For the pilot grid shown in Fig. 8 and Fig. 9 both schemes type I and type II can be applied. However, this is not generally the case. In order to motivate this problem consider the following grid

$$\mathbf{G} = \begin{bmatrix} 4 & 1 \\ 0 & 3 \end{bmatrix}$$

shown in Fig. 8. For such a grid structure it is more appropriate to employ 2×1D-PACE type I, since the pilots in frequency direction are placed along a line. On the other hand, successive pilots in time direction are shifted one subcarrier apart, which makes it difficult to employ 2×1D-PACE type II.

Accordingly, choosing the grid shown in Fig. 9, which is described by

$$\mathbf{G}_{DVB} = \begin{bmatrix} 3 & 0 \\ 1 & 4 \end{bmatrix}$$

to employ 2×1D-PACE type II is more appropriate. The grid according to Fig. 9 has been chosen for the DVB-T standard, so for DVB-T channel estimation according to a preferred embodiment of the present invention in time direction should be performed first if 2×1D-PACE is to be used.

Applications of 2×1D-PACE type I

To describe pilot symbol-assisted channel estimation according to the present

invention it is proposed to define a subset of the received signal sequence containing only the pilots, $\{\tilde{Y}^{(\mu)}_{\tilde{\ell},\tilde{i}}\} = \{Y^{(\mu)}_{\ell,i}\}$, with $\{i,\ell\} \in \mathcal{G}$, sampled at a D_f times lower rate $\tilde{i} = \lfloor i/D_f \rfloor$ in frequency direction, and at a D_t times lower rate $\tilde{\ell} = \lfloor \ell/D_t \rfloor$ in time direction, respectively.

In the following two channel estimation techniques are described for $2\times$1D-PACE type I, the first is to estimate the channel in the frequency domain by Wiener filter interpolation. The second approach is to transfer the received signal into the time domain.

Frequency Domain Channel Estimation

First Stage Wiener Filtering

For $2\times$1D-PACE type I, the first step is to estimate the channel in the frequency direction. In vector notation, the received pilot sequence of OFDM symbol ℓ becomes

$$\tilde{\mathbf{Y}}'_\ell = \sum_{\mu_1=1}^{N_{T1}} \tilde{\mathbf{X}}_1^{(\mu_1)} \tilde{\mathbf{Z}}'^{(\mu_1)}_\ell + \tilde{\mathbf{N}}'_\ell \qquad \in \mathbb{C}^{N'_P \times 1} \tag{18}$$

where $\tilde{\mathbf{Z}}'^{(\mu_1)}_\ell$ represents the received pilot sequence of subset $\mathcal{A}_{\mu_1}$ whose entries are defined in equation (14).

Further, the transmitted pilot sequence, the received pilot sequence of subset $\mathcal{A}_{\mu_1}$, and the additive noise term, of OFDM symbol ℓ transmitted from antenna μ_1, are given by

$$\begin{aligned}
\tilde{\mathbf{X}}_1^{(\mu_1)} &= \mathrm{diag}\left(\tilde{X}^{(\mu_1)}_{1_1}, \cdots, \tilde{X}^{(\mu_1)}_{1_{N'_P}}\right) \in \mathbb{C}^{N'_P \times N'_P} \\
\tilde{\mathbf{Z}}'^{(\mu_1)}_\ell &= \left[\tilde{Z}^{(\mu_1)}_{\ell,1}, \cdots, \tilde{Z}^{(\mu_1)}_{\ell,N'_P}\right]^T \in \mathbb{C}^{N'_P \times 1} \\
\tilde{\mathbf{N}}'_\ell &= \left[\tilde{N}_{\ell,1}, \cdots, \tilde{N}_{\ell,N'_P}\right]^T \in \mathbb{C}^{N'_P \times 1}
\end{aligned}$$

Further, according to the present invention channel estimation in the frequency domain is preferably performed with an FIR interpolation filter, which can be expressed for the first stage in frequency direction

$$\begin{aligned}\widehat{Z}_{\ell,i}^{(\mu_1)} &= \sum_{\bar{i}=1}^{N'_P} W'^{(\mu_1)}_{\bar{i}}[i]\, \tilde{Y}_{\bar{\ell},\bar{i}} = \sum_{\bar{i}=1}^{N'_P} W'^{(\mu_1)}_{\bar{i}}[i] \sum_{\mu_1=1}^{N_{T1}} \tilde{X}_{2\bar{\ell}}^{(\mu_2)} \tilde{Z}_{\bar{\ell},\bar{i}}^{(\mu_1)} \\ &= \mathbf{W}'^{(\mu_1)}[i]\, \tilde{\mathbf{Y}}'_{\bar{\ell}} \end{aligned} \tag{19}$$

It should be noted that in general, the filter $W'^{(\mu_1)}_{\bar{i}}[i]$ depends on the location of the desired symbol, i.e. the subcarrier index i. This means that not only for every transmit antenna but also for every subcarrier a different filter is required.

The optimum approach to estimate $\widehat{Z}_{\ell,i}^{(\mu_1)}$ is to use a Wiener interpolation filter for $\mathbf{W}'^{(\mu_1)}[i]$. A Wiener filter minimizes the means squard error MSE between the pilots sequence and the desired response. It is also known as the minimum MSE or equivalently MMSE estimator.

Further, in order to generate a Wiener filter knowledge about the channel statistics are required, which are described by the covariance matrix. The covariance matrix of the pilots in frequency direction is defined by $\mathbf{R}'_{\tilde{\mathbf{Y}}\tilde{\mathbf{Y}}} = E[\tilde{\mathbf{Y}}'_{\bar{\ell}}\ \tilde{\mathbf{Y}}'^{H}_{\bar{\ell}}]$. The entry of the m^{th} row and n^{th} column of the covariance matrix is given by

$$\{\mathbf{R}'_{\tilde{\mathbf{Y}}\tilde{\mathbf{Y}}}\}_{m,n} = E\left[\tilde{Y}_{\bar{\ell},\bar{i}-n}\, \tilde{Y}^*_{\bar{\ell},\bar{i}-m}\right] = E\left[Y_{\ell,i-D_f n}\, Y^*_{\ell,i-D_f m}\right] \qquad \{\ell, i\} \in \mathcal{G} \tag{20}$$

Further, define the cross correlation vector between the frequency response of the desired sample $Z_{\ell,i}^{(\mu_1)}$ and the pilots $\tilde{\mathbf{Y}}'_{\bar{\ell}}$. The m^{th} entry of the cross correlation vector $\mathbf{R}'^{(\mu_1)}_{Z\tilde{\mathbf{Y}}}[i] = E[Z_{\ell,i}^{(\mu_1)}\ \tilde{\mathbf{Y}}'^{H}_{\bar{\ell}}]$ can be expressed as

$$\{\mathbf{R}'^{(\mu_1)}_{Z\tilde{\mathbf{Y}}}[i]\}_m = E\left[Z_{\ell,i}^{(\mu_1)}\, \tilde{Y}^*_{\bar{\ell},\bar{i}-m}\right] = E\left[Z_{\ell,i}^{(\mu_1)}\, Y^*_{\ell,(\bar{i}-m)D_f}\right]\ ; \qquad \bar{i} = \lfloor i/D_f \rfloor \tag{21}$$

The quantities according to equation (20) and equation (21) are necessary to

evaluate the Wiener interpolation filter. The optimum solution in the MMSE sense may be determined using the Wiener-Hopf according to

$$\mathbf{W}'^{(\mu_1)}[i] = \mathbf{R}'^{(\mu_1)}_{Z\check{\mathbf{Y}}}[i] \cdot {\mathbf{R}'_{\check{\mathbf{Y}}\check{\mathbf{Y}}}}^{-1} \tag{22}$$

Second Stage Wiener Filtering

In view of the above, performing equation (19) for the N_{T1} subsets and for each sucarrier is processed further in the second stage. Here, filtering and interpolation in time direction yields the frequency response estimate which is achieved in the form

$$\widehat{H}^{(\mu)}_{\ell,i} = \sum_{\tilde{\ell}=1}^{N''_P} w''^{(\mu)}_{\tilde{\ell}}[\ell, \Delta\ell]\, \widehat{Z}^{(\mu_1)}_{\tilde{\ell},i}\,, \quad \mu = \mu_2 + N_{T2} \cdot (\mu_1 - 1) = \{1, \cdots, N_T\} \tag{23}$$

where $w''^{(\mu)}_{\tilde{\ell}}[\ell, \Delta\ell]$ represents the FIR interpolation filter of the second stage of OFDM symbol ℓ with filter delay $\Delta\ell = D_t\Delta\tilde{\ell}$.

A positive $\Delta\tilde{\ell}$ imposes a time delay of $\Delta\ell$ symbols at the receiver output. Then the estimation filter is a smoothing type filter. On the other hand, setting $\Delta\tilde{\ell} = 0$ specifies a linear prediction receiver without an induced time delay due to channel estimation, at the expense of a somewhat poorer estimate of the CIR.

It should be noted that the best performance is generally achieved if $\Delta\tilde{\ell} = N''_P/2$, i.e. the symbol to be estimated is in the middle of the pilot sequence which is used for estimation of that symbol. Therefore, for $\Delta\tilde{\ell} = N''_P/2$ there are $N''_P/2$ future and past pilot symbols involved.

Further, in matrix notation equation (23) becomes

$$\widehat{H}^{(\mu)}_{\ell,i} = \mathbf{w}''^{(\mu)}[\ell, \Delta\ell]\ \widehat{\mathbf{Z}}''^{(\mu_1)}_{i} \quad \mu = \mu_2 + N_{T2} \cdot (\mu_1 - 1) = \{1, \cdots, N_T\} \tag{24}$$

where $\mathbf{w}''^{(\mu)}[\ell, \Delta\ell] = \left[w_1''^{(\mu)}[\ell, \Delta\ell], \cdots, w_{N_P''}''^{(\mu)}[\ell, \Delta\ell]\right]$ is the channel estimation filter of stage two, and $\widehat{\mathbf{Z}}_i''^{(\mu_1)} = \left[\widehat{Z}_{1,i}^{(\mu_1)}, \cdots, \widehat{Z}_{N_P'',i}^{(\mu_1)}\right]^T$ denotes the block N_P'' outputs of stage one of subcarrier i.

In analogy to the first estimation stage, the optimum approach to estimate $\widehat{H}_{\ell,i}^{(\mu)}$ is to use a Wiener interpolation filter for $\mathbf{w}''^{(\mu)}[\ell, \Delta\ell]$. The covariance matrix of the first stage outputs in time direction is defined by $\mathbf{R}_{\widehat{\mathbf{Z}}\widehat{\mathbf{Z}}}''^{(\mu_1)} = E[\,\widehat{\mathbf{Z}}_i''^{(\mu_1)}\ \widehat{\mathbf{Z}}_i''^{(\mu_1)H}]$. The entry of the m^{th} row and n^{th} column of the covariance matrix is given by

$$\left\{\mathbf{R}_{\widehat{\mathbf{Z}}\widehat{\mathbf{Z}}}''^{(\mu_1)}\right\}_{m,n} = E\left[\widehat{Z}_{\tilde{\ell}-m,i}^{(\mu_1)}\ \widehat{Z}_{\tilde{\ell}-n,i}^{(\mu_1)\ *}\right] \tag{25}$$

Further, define the cross correlation vector between the frequency response of the desired sample $H_{\ell,i}^{(\mu)}$ and the outputs of the first stage $\widehat{\mathbf{Z}}_i''^{(\mu_1)}$. The m^{th} entry of the cross correlation vector $\mathbf{R}_{H\widehat{\mathbf{Z}}}''^{(\mu)}[\ell, \Delta\ell] = E[H_{\ell-\Delta\ell,i}^{(\mu)}\ \widehat{\mathbf{Z}}_i''^{(\mu_1)H}]$ can be expressed as

$$\left\{\mathbf{R}_{H\widehat{\mathbf{Z}}}''^{(\mu)}[\ell, \Delta\ell]\right\}_m = E\left[H_{\ell,i}^{(\mu)}\ \widehat{Z}_{\tilde{\ell}-m,i}^{(\mu_1)\ *}\right]\ ; \qquad \tilde{\ell} = \lfloor \ell/D_t \rfloor + \Delta\tilde{\ell} \tag{26}$$

The quantities in equation (20) and equation (21) are necessary to evaluate the Wiener interpolation filter. The optimum solution in the MMSE sense is derived using the Wiener-Hopf equation according to

$$\mathbf{w}''^{(\mu)}[\ell, \Delta\ell] = \mathbf{R}_{H\widehat{\mathbf{Z}}}''^{(\mu)}[\ell, \Delta\ell] \cdot \mathbf{R}_{\widehat{\mathbf{Z}}\widehat{\mathbf{Z}}}''^{(\mu_1)-1} \tag{27}$$

Time Domain Channel Estimation

Fig. 11 shows a schematic diagram of an estimator stage adapted to achieve channel estimation in the time domain according to the present invention.

As shwon in Fig. 11, an alternative approach to determine the frequency response $H_{\ell,i}$ according to the present invention is to estimate the channel impulse response (CIR), $h_{\ell,n}'^{(\mu)}$ in the time domain first. In terms of the CIR vector of OFDM

symbol ℓ impinging from the μ^{th} transmit antenna, $\tilde{\mathbf{h}}_\ell'^{(\mu)} = [\tilde{h}'^{(\mu)}_{\ell,1}, \cdots, \tilde{h}'^{(\mu)}_{\ell,Q}]^T$, the received pilot sequence of subset $\mathcal{A}_{\mu_1}$ becomes

$$\tilde{\mathbf{Z}}_\ell'^{(\mu_1)} = \sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_\ell}^{(\mu_2)} \tilde{\mathbf{F}} \tilde{\mathbf{h}}_\ell'^{(\mu)} \qquad \mu = \mu_2 + N_{T2} \cdot (\mu_1 - 1) \tag{28}$$

where $\tilde{\mathbf{F}}$ denotes an N_P'–point DFT matrix defined by

$$\{\tilde{\mathbf{F}}\}_{n,i} = e^{-j2\pi\, ni/N_P'} \,; \qquad 0 \leq i \leq N_P' - 1\,, \quad 0 \leq n \leq Q - 1 \tag{29}$$

It is assumed that the CIR is time limited to $Q \leq N_{GI}$ samples. Strictly speaking this is only true for a sample spaced channel model explained above. For a non-sample spaced channel model where the DFT of $H_{\ell,i}$ is not time limited, so oversampling is required in order to avoid aliasing. In this case, Q accounts for the number of significant taps.

Transforming $\tilde{\mathbf{Z}}_\ell'^{(\mu_1)}$ into the time domain by an N_P'–point IDFT yields

$$\begin{aligned}\tilde{\mathbf{z}}_\ell'^{(\mu_1)} \triangleq \tilde{\mathbf{F}}^H \tilde{\mathbf{Z}}_\ell'^{(\mu_1)} &= \tilde{\mathbf{F}}^H \sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_\ell}^{(\mu_2)} \tilde{\mathbf{F}} \tilde{\mathbf{h}}_\ell'^{(\mu)} \qquad \mu = \mu_2 + N_{T2} \cdot (\mu_1 - 1) \\ &= \sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_\ell}^{(\mu_2)} \tilde{\mathbf{h}}_\ell'^{(\mu)} \end{aligned} \tag{30}$$

This implies that two stage channel estimation for OFDM-based MIMO systems can be employed in the time and frequency domains in the same way.

The received pilot sequency after OFDM demodulation is given by

$$\tilde{\mathbf{Y}}_\ell' = \tilde{\mathbf{X}}_1 \tilde{\mathbf{Z}}_\ell' + \tilde{\mathbf{N}}_\ell' = \tilde{\mathbf{X}}_1 \tilde{\mathbf{F}}_{N_{T1}} \tilde{\mathbf{z}}_\ell' + \tilde{\mathbf{N}}_\ell' \tag{31}$$

where

$$\begin{aligned}
\tilde{\mathbf{X}}_1 &= \left[\tilde{\mathbf{X}}_1^{(1)}, \cdots, \tilde{\mathbf{X}}_1^{(N_{T1})}\right] \in \mathbb{C}^{N'_P \times N_{T1} N'_P} \\
\tilde{\mathbf{Z}}'_\ell &= \left[\tilde{\mathbf{Z}}'^{(1)}_\ell, \cdots, \tilde{\mathbf{Z}}'^{(N_{T1})}_\ell\right]^T \in \mathbb{C}^{N_{T1} N'_P \times 1} \\
\tilde{\mathbf{z}}'_\ell &= \left[\tilde{\mathbf{z}}'^{(1)}_\ell, \cdots, \tilde{\mathbf{z}}'^{(N_{T1})}_\ell\right]^T \in \mathbb{C}^{N_{T1} Q \times 1} \\
\tilde{\mathbf{F}}_{N_{T1}} &= \mathrm{diag}\left(\tilde{\mathbf{F}}, \cdots, \tilde{\mathbf{F}}\right) \in \mathbb{C}^{N_{T1} N'_P \times N_{T1} Q}
\end{aligned}$$

In order to estimate $\tilde{\mathbf{z}}'^{(\mu_1)}_\ell$ according to the present invention the received pilot sequency after OFDM demodulation is transformed into the time domain.

Further, for time domain channel estimation we choose to pre-multiply $\tilde{\mathbf{Y}}'_\ell$ by the transmitted pilot sequence $\tilde{\mathbf{X}}_1$ and then to transform the result into the time domain via an N'_P–point IDFT, that is

$$\begin{aligned}
\xi_\ell &\triangleq \left(\tilde{\mathbf{X}}_1 \tilde{\mathbf{F}}_{N_{T1}}\right)^H \tilde{\mathbf{Y}}'_\ell = \tilde{\mathbf{D}}'^H_\ell \tilde{\mathbf{Y}}'_\ell \qquad \in \mathbb{C}^{N_{T1} Q \times 1} \\
&= \tilde{\mathbf{D}}'^H_\ell \tilde{\mathbf{D}}'_\ell \mathbf{z}'_\ell + \tilde{\mathbf{D}}'^H_\ell \tilde{\mathbf{N}}'_\ell \qquad (32)
\end{aligned}$$

where the definition $\tilde{\mathbf{D}}'_\ell = \tilde{\mathbf{X}}_1 \tilde{\mathbf{F}}_{N_{T1}}$ has been introduced.

In the following two basic estimator structures to determine $\widehat{\mathbf{z}'}^{(\mu_1)}$ according to the present invention will be described, namely the least squares LS estimator and the minimum mean squared error MMSE estimator.

Least Squares LS Estimator

Provided the the inverse of $\mathbf{D}'^H_{\tilde{\ell}}\, \mathbf{D}'_{\tilde{\ell}}$ does exists, the least squares (LS) estimator may be determined according to

$$\begin{aligned}
\widehat{\mathbf{z}'}_{\mathrm{LS}_{\tilde{\ell}}} &= \left(\mathbf{D}'^H_{\tilde{\ell}}\, \mathbf{D}'_{\tilde{\ell}}\right)^{-1} \xi'_{\tilde{\ell}} \qquad (33) \\
&= \left(\mathbf{D}'^H_{\tilde{\ell}}\, \mathbf{D}'_{\tilde{\ell}}\right)^{-1} \mathbf{D}'^H_{\tilde{\ell}}\, \tilde{\mathbf{Y}}'_{\tilde{\ell}}
\end{aligned}$$

Since the estimator depends on the transmitted signal, the pilot sequence should be properly chosen. The LS estimator exists if $\mathbf{D}_{\tilde{\ell}}$ is full rank, unfortunately this is not always the case. A necessary condition for the LS estimator to exist is

$$N'_P \geq N_{T1}\,Q \tag{34}$$

According to a preferred embodiment of the present invention, two times over-sampling provides a good trade-off between minimizing the system overhead due to pilots and optimizing the performance, i.e. $N'_P \approx 2N_{T1}\,Q$. It is assumed that $N_{GI} \geq Q$, i.e. the guard interval is longer than the maximum delay of the channel.

Further, it should be noted that the LS estimator for more than one transmit antenna does only exist in the time domain.

Minimum Mean Squared Error MMSE Estimator

The MMSE estimator is given by an FIR filter which is for time domain channel estimation

$$\begin{aligned} \hat{z}^{(\mu_1)}_{\tilde{\ell},n} &= \sum_{m=1}^{N'_P} w'^{(\mu_1)}_{m}[n]\,\xi_{\tilde{\ell},m} = \sum_{m=1}^{N'_P} w'^{(\mu_1)}_{i,m}[n] \sum_{\mu_1=1}^{N_{T1}} \tilde{X}^{(\mu_2)}_{2_{\tilde{\ell}}}\,\tilde{z}^{(\mu_1)}_{\tilde{\ell},m} \\ &= \mathbf{w}'^{(\mu_1)}[n]\,\boldsymbol{\xi}'_{\tilde{\ell}} \qquad n = \{1,\cdots,N_c\} \end{aligned} \tag{35}$$

The Wiener filter is determined by the Wiener-Hopf equation

$$\mathbf{w}'^{(\mu_1)}[n] = \mathbf{R}'^{(\mu_1)}_{z\xi}[n] \cdot \mathbf{R}'^{-1}_{\xi\xi} \qquad n = \{1,\cdots,N_c\} \tag{36}$$

In general, the Wiener filter $\mathbf{w}'^{(\mu_1)}[n]$, depends on the location of the desired symbol n. In order to generate the MMSE estimator knowledge of the correlation

matrices $\mathbf{R}'_{\xi\xi}$ and $\mathbf{R}'^{(\mu_1)}_{z\xi}[n]$ are required

$$\begin{aligned}\mathbf{R}'_{\xi\xi} &\triangleq E\left\{\xi'_{\tilde{\ell}}\,\xi'^{H}_{\tilde{\ell}}\right\} = \mathbf{D}'^{H}_{\tilde{\ell}}\,\mathbf{R}'_{\tilde{\mathbf{Y}}\tilde{\mathbf{Y}}}\,\mathbf{D}'_{\tilde{\ell}} \quad \in \mathbb{C}^{QN_{T1}\times QN_{T1}} \\ &= \mathbf{D}'^{H}_{\tilde{\ell}}\,\tilde{\mathbf{X}}'_{\tilde{\ell}}\,\mathbf{R}'_{\tilde{\mathbf{Z}}\tilde{\mathbf{Z}}}\,\tilde{\mathbf{X}}'^{H}_{\tilde{\ell}}\,\mathbf{D}'_{\tilde{\ell}} + N_0\,\mathbf{D}'^{H}_{\tilde{\ell}}\,\mathbf{D}'_{\tilde{\ell}} \\ &= \mathbf{D}'^{H}_{\tilde{\ell}}\,\mathbf{D}'_{\tilde{\ell}}\,\mathbf{R}'_{\tilde{\mathbf{z}}\tilde{\mathbf{z}}}\,\mathbf{D}'^{H}_{\tilde{\ell}}\,\mathbf{D}'_{\tilde{\ell}} + N_0\,\mathbf{D}'^{H}_{\tilde{\ell}}\,\mathbf{D}'_{\tilde{\ell}}\end{aligned} \tag{37}$$

and

$$\begin{aligned}\mathbf{R}'^{(\mu_1)}_{z\xi}[n] &\triangleq E\left\{z^{(\mu_1)}_{\tilde{\ell},n}\,\xi'^{H}_{\tilde{\ell}}\right\} \quad \in \mathbb{C}^{1\times QN_{T1}} \\ &= \mathbf{R}'^{(\mu_1)}_{z\tilde{\mathbf{z}}}[n]\,\mathbf{D}'^{H}_{\tilde{\ell}}\,\mathbf{D}'_{\tilde{\ell}}\end{aligned} \tag{38}$$

Further, the covariance matrix of $\tilde{\mathbf{z}}'_{\tilde{\ell}}$ is denoted by $\mathbf{R}'_{\tilde{\mathbf{z}}\tilde{\mathbf{z}}} = E\{\tilde{\mathbf{z}}'_{\tilde{\ell}}\,\tilde{\mathbf{z}}'^{H}_{\tilde{\ell}}\}$ with dimension $QN_{T1}\times QN_{T1}$. Furtermore, $\mathbf{R}'^{(\mu_1)}_{z\tilde{\mathbf{z}}}[n]$ is row $n+(\mu_1-1)Q$ of $\mathbf{R}'_{\tilde{\mathbf{z}}\tilde{\mathbf{z}}}$. The covariance matrix in the time domain $\mathbf{R}'_{\tilde{\mathbf{z}}\tilde{\mathbf{z}}}$ is related to the covariance matrix in the frequency domain $\mathbf{R}_{\tilde{\mathbf{Z}}\tilde{\mathbf{Z}}}$ by

$$\mathbf{R}'_{\tilde{\mathbf{z}}\tilde{\mathbf{z}}} = \tilde{\mathbf{F}}^{H}_{N_{T1}}\,\mathbf{R}'_{\tilde{\mathbf{Z}}\tilde{\mathbf{Z}}}\,\tilde{\mathbf{F}}_{N_{T1}} \tag{39}$$

Here, it is assumed that the fading of different transmit antennas is uncorrelated. It should be noted that while the LS estimator requires $\mathbf{D}'_{\tilde{\ell}}$ to be full rank, while the MMSE estimator requires invertability of $\mathbf{R}'_{\xi\xi}$ as seen from equation (36). For this to hold, however, $\mathbf{D}'_{\tilde{\ell}}$ need not to be full rank. Thus, the MMSE estimator can exist even if $N'_P < N_{T1}\,Q$.

For the case that $\mathbf{D}'_{\tilde{\ell}}$ is full rank, the inverse of $\mathbf{D}'^{H}_{\tilde{\ell}}\,\mathbf{D}'_{\tilde{\ell}}$ does exist. Then the Wiener filter of equation (36) can be simplied to

$$\begin{aligned}\mathbf{w}'^{(\mu_1)}[n] &= \mathbf{R}'^{(\mu_1)}_{z\tilde{\mathbf{z}}}[n]\cdot\left[\mathbf{R}'_{\tilde{\mathbf{z}}\tilde{\mathbf{z}}} + \left(\mathbf{D}'^{H}_{\tilde{\ell}}\,\mathbf{D}'_{\tilde{\ell}}\right)^{-1} N_0\right]^{-1}\cdot\left(\mathbf{D}'^{H}_{\tilde{\ell}}\,\mathbf{D}'_{\tilde{\ell}}\right)^{-1} \\ \hat{z}^{(\mu_1)}_{\tilde{\ell},n} &= \mathbf{R}'^{(\mu_1)}_{z\tilde{\mathbf{z}}}[n]\cdot\left[\mathbf{R}'_{\tilde{\mathbf{z}}\tilde{\mathbf{z}}} + \left(\mathbf{D}'^{H}_{\tilde{\ell}}\,\mathbf{D}'_{\tilde{\ell}}\right)^{-1} N_0\right]^{-1}\cdot\hat{z}^{(\mu_1)}_{LS\,\tilde{\ell},n}\end{aligned} \tag{40}$$

Further it should be noted that according to the present invention the separation

of the N_{T1} signals, which is performed by the LS estimator, can be separated from the filtering task.

As outlined above, the MMSE estimator is in general dependent on the choice of the pilot symbols. However, choosing orthogonal pilot sequences $\tilde{\mathbf{X}}_1^{(\mu_1)}$ the estimator becomes independent on the transmitted pilots. For orthogonal pilots $\tilde{\mathbf{X}}_1^{(\mu)} \tilde{\mathbf{X}}_1^{(m)^T} = \mathbf{I} \cdot \delta_{\mu,m}$, where $\delta_{\mu,m}$ denotes the Kronecker symbol, it can be shown that $\mathbf{D'}_{\tilde{\ell}}^H \mathbf{D'}_{\tilde{\ell}} = N'_P\mathbf{I}$.

Therefore, the above LS estimator in equation (33) as well as the MMSE estimator in equation (40) can be grossly simplified, since the matrix inversion required in equation (33) and equation (40) become straightforward.

Further, it can be seen from equation (33) and equation (40) that the estimator has become independent of the chosen pilot sequence, which does significantly simplify the filter generation.

Fig. 11 shows a block diagram of channel estimation and interpolation in the time domain using orthogonal pilot sequences. First, the received pilot sequence is split into N_{T1} branches and each branch pre-multiplied by $\tilde{\mathbf{X}}_1^{(\mu_1)}$. Then each branch of the received pilot sequence is transformed to the time domain. By means of – optional – filtering and/or windowing the channel is estimated in the time domain. Zero padding of the first stage estimate extends its lenght to N_c samples. The estimate of the CTF of an entire OFDM symbol (pilots and data), is obtained by an N_c–point FFT of the CIR estimate

$$\widehat{\mathbf{Z}}' = \mathbf{F}_{N_{T1}} \widehat{\mathbf{z}}' \qquad \text{or} \qquad \widehat{\mathbf{Z}}'^{(\mu_1)} = \mathbf{F}\ \widehat{\mathbf{z}}'^{(\mu_1)} \tag{41}$$

where $\mathbf{F}_{N_{T1}}$ is a $N_{T1}N_c \times N_{T1}N_c$ block diagonal matrix, consisting of N_{T1} blocks of N_c-point DFT matrices $\mathbf{F}$.

As shown in Fig. 11, the output of the first stage $\widehat{\mathbf{Z}}'^{(\mu_1)}$ can be fed to the second

stage estimator in equation (23).

Alternatively, $\widehat{\mathbf{z}}'^{(\mu_1)}$ may be fed into (23) to yield the CIR estimate $\widehat{h}_{\ell,i}^{(\mu)}$ which is then transformed into the frequency domain with N_{T1} FFTs. Moreover, channel estimation of the second stage in the Doppler domain is possible, that is $\widehat{\mathbf{Z}}'^{(\mu_1)}$ or $\widehat{\mathbf{z}}'^{(\mu_1)}$ are transformed into the Doppler domain using equivalent algorithms as in the time domain.

Fig. 12 shows a schematic diagram of an estimator stage adapted to achieve channel estimation in the time domain according to the present invention.

From Fig. 12 it can be seen that the channel estimation of the second stage may also be performed with DFT-interpolation cooresponding to the first estimation stage discribed above. However, it may be computationally more efficient to perform the second estimation stage in the time domain as well, i.e., before zero padding and the N_c-point FFT.

As shown in Fig. 12, the filtering of the second stage itself remains unaffected as the correlation function in frequency and time – i.e. the first and second dimension of transmission – are mutually independent. Therefore, of Wiener filtering is chosen for the second stagem equations (23) and (27) referred to above may still be used, the only difference being that the input becomes $\hat{z}_n''^{(\mu_1)}$ instead of $\hat{Z}_i''^{(\mu_1)}$, and the output is $\hat{h}_{l,n}^{(\mu)}$ instead of $\hat{H}_{l,n}^{(\mu)}$, where n is the sample index in the time domain. After the second stage filtering each of the overall N_T outputs – i.e., N_{T2} outputs per subset – are transformed back into the frequency domain.

Application to a Cellular System with a Frequency Reuse Factor of One

Fig. 13 shows an aplplication of the two stage channnel estimation approach according to the present invention to a cellular communication system with a frequency reuse factor of one.

As shown in Fig. 13, instead of having an antenna array with N_T antenna

elements, the proposed scheme can be applied to distributed antennas as well. E.g., an application is to employ 2×1D-PACE to a celluar system with a frequency reuse factor of one. In a scenario where the mobile user is at the cell border, the user will receive the desired signal from one base station and one or several interfering signals from other base stations.

Further, one may assume that each base station has N_{T2} antenna elements. While the data bearing symbols can be protected against interference using a channel code or by spreading, the pilot symbols cannot be protected in this way. Accurate channel estimation, however, is most important for the system to work efficiently. One solution is to boost the pilots; this however will increase the interference to users served by other base stations, and thus limits the system capcity.

According to the present invention, 2×1D-PACE can be applied to this scenario as follows: the base stations form N_{T1} subsets, each subset having an antenna array with N_{T2} antenna elements, to form an resulting array of $N_T = N_{T1}N_{T2}$ elements. This would require inter-cell synchronization.

Abbreviations

AWGN Additive white Gaussian noise

CIR Channel impulse response

CTF Channel transfer function

DFT Discrete Fourier transform

FFT Fast Fourier transform

GI Guard interval

IDFT Inverse discrete Fourier transform

IFFT Inverse fast Fourier transform

LS Least squares

MIMO Multiple input multiple output, generally a system having several transmit and receive antennas

MMSE Minimum mean squared error

MSE Mean squared error

OFDM Orthogonal frequency division multiplexing

PACE Pilot-symbol aided channel estimation

List of Commonly Used System Parameters

N_{FFT} FFT length.

N_c Number of subcarriers.

N_{GI} Number of samples of the guard interval.

L Number of OFDM symbols per frame.

T OFDM symbol duration.

T_{spl} Sample interval, given by $T_{spl} = T/N_{FFT}$.

T_{sym} Total OFDM symbol duration including the guard interval $T_{spl} = T + N_{GI}T_{spl}$.

Q_0 Number of non-zero channel taps.

Q Total number of channel taps.

N_R Number of receive antennas.

N_T Number of transmit antennas.

N_{T1} Number of subsets.

N_{T2} Number of transmit antennas within one subset.

N_P' Number of pilots in frequency direction.

N_P'' Number of pilots in time direction.

N_P Number of total pilots used for channel estimation ($N_P = N_P' N_P''$).

D_f Pilot spacing in frequency.

D_t Pilot spacing in time.

List of Commonly Used Variables

$X_{\ell,i}^{(\mu)}$ Transmitted OFDM symbol of the μ^{th} transmit antenna of OFDM symbol ℓ at subcarrier i.

$x^{(\mu)}(t)$ Transmitted signal of the μ^{th} transmit antenna after OFDM modulation.

$Y_{\ell,i}^{(\nu)}$ Received OFDM symbol of the ν^{th} receive antenna of OFDM symbol ℓ at subcarrier i.

$y^{(\nu)}(t)$ Received signal of the ν^{th} receive antenna at time t before OFDM demodulation.

$H_{\ell,i}^{(\mu,\nu)}$ Channel transfer function (CTF) of the ν^{th} receive antenna arriving from the μ^{th} transmit antenna of OFDM symbol ℓ at subcarrier i.

$h^{(\mu,\nu)}(t)$ Channel impulse response (CIR) of the ν^{th} receive antenna arriving from the μ^{th} transmit antenna.

$h_{\ell,n}^{(\mu,\nu)}$ Sampled CIR of the ν^{th} receive antenna arriving from the μ^{th} transmit antenna, at the n^{th} sample of OFDM symbol ℓ.

$N_{\ell,i}$ Sample of AWGN with zero mean and variance N_0 of OFDM symbol ℓ at subcarrier i.

$n(t)$ Realization of a AWGN process at time t before OFDM demodulation.

Claims

1. Method of two dimensional channel estimation for multiple input multiple output transmission systems using multicarrier modulated transmission signals impinging from a plurality of (N_T) transmit antennas and carrying a two dimensional data sequence with embedded pilot symbols, comprising the steps:

dividing the plurality of transmit antennas into disjoint transmission antenna subsets ($\mathcal{A}_{\mu_1}$);

separating impinging pilot sequences in relation to transmission antenna subsets ($\mathcal{A}_{\mu_1}$) by performing a first stage channel estimation to yield tentative estimates of a channel response in a first dimension of transmission;

seperating impinging pilot sequences in relation to antennas in transmission antenna subsets ($\mathcal{A}_{\mu_1}$) by performing a second stage channel estimation for each antenna in each transmission antenna subset ($\mathcal{A}_{\mu_1}$) to yield an estimation of the channel response.

2. Method according to claim 1, **characterized in that** the first stage channel estimation is performed using pilot sequences arranged as a two dimensional grid of pilot symbols, wherein pilot symbols used for first stage channel estimation depend on the first dimension of transmission only and pilot symbols used for second stage channel estimation depend on a second dimension of trans-

mission only.

3. Method according to claim 2, **characterized in that** pilot sequences are expressed in a product form for achieving seperability of pilot sequences in the first dimension of transmission and the second dimension of transmission.

4. Method according to claim 3, **characterized in that** pilot sequences are selected from orthogonal designs.

5. Method according to claim 4, **characterized in that** orthogonal designs are selected from a group comprising Walsh sequences and phase shifted sequences.

6. Method according to one of the claims 1 to 5, **characterized in that** the first dimension is frequency dimension.

7. Method according to claim 6, **characterized in that** pilot sequences are expressed in the product form according to

$$\tilde{X}^{(\mu)}_{\tilde{\ell},\tilde{i}} = \tilde{X}^{(\mu_1)}_{1_{\tilde{i}}} \cdot \tilde{X}^{(\mu_2)}_{2_{\tilde{\ell}}}, \qquad \begin{aligned} \mu &= \mu_2 + N_{T2} \cdot (\mu_1 - 1) \\ \mu_1 &= \{1, \cdots, N_{T1}\} \\ \mu_2 &= \{1, \cdots, N_{T2}\} \end{aligned}$$

wherein

$\tilde{l}$ referes to a vector of data symbols in the two dimensional data sequence arranged in time direction;

$\tilde{i}$ is a sub-carrier index in frequency direction;

$\tilde{X}_{1_{\tilde{i}}}^{(\mu_1)}$ represent pilot symbols for the first stage channel estimation in the first dimension of transmission;

$\tilde{X}_{2_{\tilde{i}}}^{(\mu_2)}$ represent pilot symbols for the second stage channel estimation in a second dimension of transmission; and

N_{T1} is the number of transmission antenna subsets ($\mathcal{A}_{\mu_1}$), N_{T2} is the number of antennas in the transmission antenna subsets ($\mathcal{A}_{\mu_1}$), and a product of the number of transmission antenna subsets and the number of antennas in the transmission antenna subsets equals the number N_T of the plurality of transmit antennas $N_T = N_{T1} \cdot N_{T2}$.

8. Method according to one of the claims 6 or 7, **characterized in that** for a pilot spacing having a value of one first stage channel estimation and/or second stage channel estimation is achieved in a non-interpolating manner through yield of tentative estimates in relation to pilot symbol grid positions in the dimension of estimation.

9. Method according to one of the claims 6 or 7, **characterized in that** for a pilot spacing having a value larger than one first stage channel estimation and/or second stage channel estimation is achieved in an interpolating manner through yield of tentative estimates for all data sequence grid positions in the dimension of estimation.

10. Method according to one of the claims 7 to 9, **characterized in that** the first stage channel estimation is achieved using the product form of pilot sym-

bols and a representation of received pilot symbols according to

$$\tilde{Y}_{\tilde{\ell},\tilde{i}} = \sum_{\mu_1=1}^{N_{T1}} \tilde{X}_{1_{\tilde{i}}}^{(\mu_1)} \underbrace{\sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_{\tilde{\ell}}}^{(\mu_2)}\, \tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu_2 + N_{T2}\cdot(\mu_1 - 1))}}_{\tilde{Z}_{\tilde{\ell},\tilde{i}}^{(\mu_1)}} + \tilde{N}_{\tilde{\ell},\tilde{i}}$$

wherein

$$\tilde{Z}_{\tilde{\ell},\tilde{i}}^{(\mu_1)} \triangleq \sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_{\tilde{\ell}}}^{(\mu_2)}\, \tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu_2 + N_{T2}\cdot(\mu_1 - 1))}$$

is a representation of a received pilot sequence to be estimated in relation to the transmission antenna subset $\mathcal{A}_{\mu_1}$;

$\tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu)}$ is the frequency response of transmit antenna μ; and

$\tilde{N}_{\tilde{\ell},\tilde{i}}$ is noise superimposed during transmission.

11. Method according to one of the claims 6 to 10, **characterized in that** the first stage channel estimation is achieved in frequency domain.

12. Method according to one of the claims 6 to 10, **characterized in that** the first stage channel estimation is achieved in time domain.

13. Method according to claim 11 or 12, **characterized in that** the second stage channel estimation is achieved in frequency domain.

14. Method according to claim 11 or 12, **characterized in that** the second stage channel estimation is achieved in time domain.

15. Method according to one of the claims 1 to 5, **characterized in that**

the first dimension is time dimension.

16. Method according to claim 15, **characterized in that** pilot sequences are expressed in the product form according to

$$\tilde{X}^{(\mu)}_{\tilde{\ell},\tilde{i}} = \tilde{X}^{(\mu_1)}_{1_{\tilde{\ell}}} \cdot \tilde{X}^{(\mu_2)}_{2_{\tilde{i}}}\,, \quad \mu = \mu_2 + N_{T2} \cdot (\mu_1 - 1)\,, \quad \begin{array}{l} \mu_1 = \{1, \cdots, N_{T1}\} \\ \mu_2 = \{1, \cdots, N_{T2}\} \end{array}$$

wherein

$\tilde{l}$ referes to a vector of data symbols in the two dimensional data sequence arraged in time direction;

$\tilde{i}$ is a sub-carrier index in freqeuncy direction;

$\tilde{X}^{(\mu_1)}_{1_{\tilde{\ell}}}$ represent pilot symbols for the first stage channel estimation in the first dimension of transmission;

$\tilde{X}^{(\mu_2)}_{2_{\tilde{i}}}$ represent pilot symbols for the second stage channel estimation in a second dimension of transmission; and

N_{T1} is the number of transmission antenna subsets ($\mathcal{A}_{\mu_1}$), N_{T2} is the number of antennas in the transmission antenna subsets ($\mathcal{A}_{\mu_1}$), and a product of the number of transmission antenna subsets and the number of antennas in the transmission antenna subsets equals the number N_T of the plurality of transmit antennas $N_T = N_{T1} \cdot N_{T2}$.

17. Method according to one of the claims 15 or 16, **characterized in that** for a pilot spacing having a value of one first stage channel estimation and/or second stage channel estimation is achieved in a non-interpolating manner through yield of tentative estimates in relation to pilot symbol grid positions in the dimen-

sion of estimation.

18. Method according to one of the claims 15 or 16, **characterized in that** for a pilot spacing having a value larger than one first stage channel estimation and/or second stage channel estimation is achieved in an interpolating manner through yield of tentative estimates for all data sequence grid positions in the dimension of estimation.

19. Method according to claim one of the claims 16 to 18, **characterized in that** the first stage channel estimation is achieved using the product form of pilot symbols and a representation of received pilot symbols according to

$$\tilde{Y}_{\bar{\ell},\bar{i}} = \sum_{\mu_1=1}^{N_{T1}} \tilde{X}_{1_{\bar{\ell}}}^{(\mu_1)} \underbrace{\sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_{\bar{i}}}^{(\mu_2)} \tilde{H}_{\bar{\ell},\bar{i}}^{(\mu_2 + N_{T2} \cdot (\mu_1 - 1))}}_{\tilde{Z}_{\bar{\ell},\bar{i}}^{(\mu_1)}} + \tilde{N}_{\bar{\ell},\bar{i}}$$

wherein

$$\tilde{Z}_{\bar{\ell},\bar{i}}^{(\mu_1)} \triangleq \sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_{\bar{i}}}^{(\mu_2)} \tilde{H}_{\bar{\ell},\bar{i}}^{(\mu_2 + N_{T2} \cdot (\mu_1 - 1))}$$

is a representation of a received pilot sequence to be estimated in relation to transmission antenna subset $\mathcal{A}_{\mu_1}$;

$\tilde{H}_{\bar{\ell},\bar{i}}^{(\mu)}$ is the frequency response of transmit antenna μ; and

$\tilde{N}_{\bar{\ell},\bar{i}}$ is noise superimposed during transmission.

20. Method according to one of the claims 15 to 19, **characterized in that** the first stage channel estimation is achieved in frequency domain.

21. Method according to one of the claims 15 to 19, **characterized in**

that the first stage channel estimation is achieved in time domain.

22. Method according to claim 20 or 21, **characterized in that** the second stage channel estimation is achieved in frequency domain.

23. Method according to claim 20 or 21, **characterized in that** the second stage channel estimation is achieved in time domain.

24. Method according to claim 11 or 20, **characterized in that** the first stage channel estimation in frequency domain is achieved through finite impulse response filtering of received pilot symbols to yield tentative estimates ($\hat{Z}^{(\mu_1)}_{\tilde{\ell},\tilde{i}}$) of received pilot sequences ($\tilde{Z}^{(\mu_1)}_{\tilde{\ell},\tilde{i}}$) in relation to transmission antenna subsets ($\mathcal{A}_{\mu_1}$).

25. Method according to claim 13 or 22, **characterized in that** the second stage channel estimation in frequency domain is achieved through finite impulse response filtering of tentative estimates ($\hat{Z}^{(\mu_1)}_{\tilde{\ell},\tilde{i}}$) in relation to antennas in transmission antenna subsets ($\mathcal{A}_{\mu_1}$) to yield estimates of the channel reponse ($\hat{H}^{(\mu)}_{\ell,i}$).

26. Method according to claim 24 or 25, **characterized in that** the second stage channel estimation in frequency domain is executed through finite impulse response filtering with induced time delay for smoothing type filtering.

27. Method according to claim 24 or 25, **characterized in that** the second stage channel estimation in frequency domain is executed through finite impulse response filtering without induced time delay for predictor type filtering.

28. Method according to one of the claims 24 to 27, **characterized in**

that the finite impulse response filtering is a Wiener type filtering.

29. Method according to claim 12 or 21, **characterized in that** the first stage channel estimation in time domain is achieved through executing an inverse fourier transformation of received pilot symbols, executing a finite impulse response filtering of the inverse fourier transformation result, zero padding, and executing fourier transformation of the result of zero padding to yield tentative estimates ($\hat{Z}_{\ell,\tilde{i}}^{(\mu_1)}$) in relation to antennas in transmission antenna subsets ($\mathcal{A}_{\mu_1}$).

30. Method according to claim 14 or 23, **characterized in that** the second stage channel estimation in time domain is achieved through executing an inverse fourier transformation of tentative estimates ($\hat{Z}_{\ell,\tilde{i}}^{(\mu_1)}$) in relation to antennas in transmission antenna subsets ($\mathcal{A}_{\mu_1}$), optionally executing a finite impulse response filtering of the inverse fourier transformation result, zero padding, and executing a fourier transformation of the result of zero padding to yield estimates of the channel reponse ($\hat{H}_{\ell,i}^{(\mu)}$).

31. Method according to claim 29 or 30, **characterized in that** it further comprises a step of interpolating the result of finite impulse filtering before fourier transformation thereof for synthesizing a number of tentative estimates in the time domain corresponding to the number of sub-carriers of the two dimensional data sequence.

32. Method according to claim 29 or 30, **characterized in that** a ratio of the number of tentative estimates in the time domain to the number of received pilot symbols corresponds to pilot spacing in the first dimension of transmission.

33. Method according to one of the claims 29 to 32, **characterized in that** the second stage channel estimation is excuted on the result of finite impulse

filtering.

34. Method according to one of the claims 1 to 33, **characterized in that** it is applied to a cellular communication system with a frequency reuse factor of one such that base stations and related anntenna arrays form the plurality of transmit antennas and such that transmission antenna subsets ($\mathcal{A}_{\mu_1}$) and related transmission antennas are defined in relation to this plurality of transmit antennas.

35. Channel estimator for estimating multiple input multiple output transmission channels in two dimensions, wherein transmission signals impinge from a plurality of (N_T) transmit antennas and carry a two dimensional data sequence with embedded pilot symbols, comprising:

a memory storing a division of the plurality of transmit antennas into disjoint transmission antenna subsets ($\mathcal{A}_{\mu_1}$);

at least one first estimator stage adapted to separate impinging pilot sequences in relation to transmission antenna subsets ($\mathcal{A}_{\mu_1}$) and to perform a first stage channel estimation to yield tentative estimates of a channel response in a first dimension of transmission;

at least one second estimator stage adapted to seperate impinging pilot sequences in relation to antennas in transmission antenna subsets ($\mathcal{A}_{\mu_1}$) and to perform a second stage channel estimation for each antenna in each transmission antenna subset ($\mathcal{A}_{\mu_1}$) to yield an estimation of the channel response.

36. Channel estimator according to claim 35, **characterized in that** the first estimator stage is adapted to perform the first stage channel estimation using pilot sequences arranged as a two dimensional grid of pilot symbols, wherein pilot symbols used for first stage channel estimation depend on the first dimension

of transmission only and pilot symbols used for second stage channel estimation depend on a second dimension of transmission only.

37. Channel estimator according to claim 36, **characterized in that** the first stage estimator stage is adapted to perform the first stage channel estimation using pilot sequences using pilot sequences expressed in a product form for achieving seperability of pilot sequences in the first dimension of transmission and the second dimension of transmission.

38. Channel estimator according to claim 37, **characterized in that** the first stage estimator stage is adapted to perform the first stage channel estimation using pilot sequences selected from orthogonal designs.

39. Channel estimator according to claim 38, **characterized in that** the first stage estimator stage is adapted to perform the first stage channel estimation using Walsh pilot sequences or phase shifted pilot sequences.

40. Channel estimator according to one of the claims 35 to 39, **characterized in that** the first estimator stage is adapted to perform the first stage channel estimation in frequency dimension.

41. Channel estimator according to claim 40, **characterized in that** the first estimator stage is adapted to perform the first stage channel estimation using pilot sequences expressed in the product form according to

$$\tilde{X}^{(\mu)}_{\tilde{\ell},\tilde{i}} = \tilde{X}^{(\mu_1)}_{1_{\tilde{i}}} \cdot \tilde{X}^{(\mu_2)}_{2_{\tilde{\ell}}}, \qquad \begin{aligned} \mu &= \mu_2 + N_{T2} \cdot (\mu_1 - 1) \\ \mu_1 &= \{1, \cdots, N_{T1}\} \\ \mu_2 &= \{1, \cdots, N_{T2}\} \end{aligned}$$

wherein

$\tilde{l}$ referes to a vector of data symbols in the two dimensional data sequence arranged in time direction;

$\tilde{\imath}$ is a sub-carrier index in freqeuncy direction;

$\tilde{X}_{1_{\tilde{\imath}}}^{(\mu_1)}$ represent pilot symbols for the first stage channel estimation in the first dimension of transmission;

$\tilde{X}_{2_{\tilde{l}}}^{(\mu_2)}$ represent pilot symbols for the second stage channel estimation in a second dimension of transmission; and

N_{T1} is the number of transmission antenna subsets ($\mathcal{A}_{\mu_1}$), N_{T2} is the number of antennas in the transmission antenna subsets ($\mathcal{A}_{\mu_1}$), and a product of the number of transmission antenna subsets and the number of antennas in the transmission antenna subsets equals the number N_T of the plurality of transmit antennas $N_T = N_{T1} \cdot N_{T2}$.

42. Channel estimator according claim 40 or 41, **characterized in that** the first stage estimator and/or the second stage estimator are adapted to achieve channel estimation for a pilot spacing having a value of one in a non-interpolating manner through yield of tentative estimates in relation to pilot symbol grid positions.

43. Method according to one of the claims 40 to 41, **characterized in that** the first stage estimator and/or the second stage estimator are adapted to achieve channel estimation for a pilot spacing having a value larger than one channel estimation in an interpolating manner through yield of tentative estimates for

all data sequence grid positions in the dimension of estimation.

44. Channel estimator according to one of the claims 41 to 43, **characterized in that** the first estimator stage is adapted to perform the first stage channel estimation using the product form of pilot symbols and a representation of received pilot symbols according to

$$\tilde{Y}_{\tilde{\ell},\tilde{i}} = \sum_{\mu_1=1}^{N_{T1}} \tilde{X}_{1_{\tilde{i}}}^{(\mu_1)} \underbrace{\sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_{\tilde{\ell}}}^{(\mu_2)} \tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu_2 + N_{T2}\cdot(\mu_1 - 1))}}_{\tilde{Z}_{\tilde{\ell},\tilde{i}}^{(\mu_1)}} + \tilde{N}_{\tilde{\ell},\tilde{i}}$$

wherein

$$\tilde{Z}_{\tilde{\ell},\tilde{i}}^{(\mu_1)} \triangleq \sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_{\tilde{\ell}}}^{(\mu_2)} \tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu_2 + N_{T2}\cdot(\mu_1 - 1))}$$

is a representation of a received pilot sequence to be estimated in relation to the transmission antenna subset $\mathcal{A}_{\mu_1}$;

$\tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu)}$ is the frequency response of transmit antenna μ; and

$\tilde{N}_{\tilde{\ell},\tilde{i}}$ is noise superimposed during transmission.

45. Channel estimator according to one of the claims 40 to 44, **characterized in that** the first estimator stage is adapted to perform the first stage channel estimation in frequency domain.

46. Channel estimator according to one of the claims 40 to 44, **characterized in that** the first estimator stage is adapted to perform the first stage channel estimation in time domain.

47. Channel estimator according to claim 45 or 46, **characterized in that**

second estimator stage is adapted to perform the second stage channel estimation in frequency domain.

48. Channel estimator according to claim 45 or 46, **characterized in that** second estimator stage is adapted to perform the second stage channel estimation in time domain.

49. Channel estimator according to one of the claims 35 to 39, **characterized in that** the first estimator stage is adapted to perform the first stage channel estimation in time dimension.

50. Channel estimator according to claim 49, **characterized in that** the first estimator stage is adapted to perform the first stage channel estimation using pilot sequences expressed in the product form according to

$$\tilde{X}^{(\mu)}_{\tilde{\ell},\tilde{i}} = \tilde{X}^{(\mu_1)}_{1_{\tilde{\ell}}} \cdot \tilde{X}^{(\mu_2)}_{2_{\tilde{i}}}\,, \quad \mu = \mu_2 + N_{T2} \cdot (\mu_1 - 1)\,, \quad \begin{array}{l} \mu_1 = \{1, \cdots, N_{T1}\} \\ \mu_2 = \{1, \cdots, N_{T2}\} \end{array}$$

wherein

$\tilde{l}$ referes to a vector of data symbols in the two dimensional data sequence arraged in time direction;

$\tilde{i}$ is a sub-carrier index in freqeuncy direction;

$\tilde{X}^{(\mu_1)}_{1_{\tilde{\ell}}}$ represent pilot symbols for the second stage channel estimation in the first dimension of transmission;

$\tilde{X}^{(\mu_2)}_{2_{\tilde{i}}}$ represent pilot symbols for the first stage channel estimation in a second dimension of transmission; and

N_{T1} is the number of transmission antenna subsets $(\mathcal{A}_{\mu_1})$, N_{T2} is the number of

antennas in the transmission antenna subsets ($\mathcal{A}_{\mu_1}$), and a product of the number of transmission antenna subsets and the number of antennas in the transmission antenna subsets equals the number N_T of the plurality of transmit antennas $N_T = N_{T1} \cdot N_{T2}$.

51. Method according to one of the claims 49 or 50, **characterized in that** for a pilot spacing having a value of one first stage channel estimation and/or second stage channel estimation is achieved in a non-interpolating manner through yield of tentative estimates in relation to pilot symbol grid positions.

52. Method according to one of the claims 49 or 50, **characterized in that** for a pilot spacing having a value larger than one first stage channel estimation and/or second stage channel estimation is achieved in an interpolating manner through yield of tentative estimates for all data sequence grid positions in the dimension of estimation.

53. Channel estimator according to one of the claims 50 to 52, **characterized in that** the first estimator stage is adapted to perform the first stage channel estimation using the product form of pilot symbols and a representation of received pilot symbols according to

$$\tilde{Y}_{\tilde{\ell},\tilde{i}} = \sum_{\mu_1=1}^{N_{T1}} \tilde{X}_{1_{\tilde{\ell}}}^{(\mu_1)} \underbrace{\sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_{\tilde{i}}}^{(\mu_2)} \tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu_2 + N_{T2} \cdot (\mu_1 - 1))}}_{\tilde{Z}_{\tilde{\ell},\tilde{i}}^{(\mu_1)}} + \tilde{N}_{\tilde{\ell},\tilde{i}}$$

wherein

$$\tilde{Z}_{\tilde{\ell},\tilde{i}}^{(\mu_1)} \triangleq \sum_{\mu_2=1}^{N_{T2}} \tilde{X}_{2_{\tilde{i}}}^{(\mu_2)} \tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu_2 + N_{T2} \cdot (\mu_1 - 1))}$$

is a representation of a received pilot sequence to be estimated in relation to transmission antenna subset $\mathcal{A}_{\mu_1}$;

$\tilde{H}_{\tilde{\ell},\tilde{i}}^{(\mu)}$ is the frequency response of transmit antenna μ; and

$\tilde{N}_{\tilde{\ell},\tilde{i}}$ is noise superimposed during transmission.

54. Channel estimator according to one of the claims 49 to 53, **characterized in that** the first estimator stage is adapted to perform the first stage channel estimation in frequency domain.

55. Channel estimator according to one of the claims 49 to 53, **characterized in that** the first estimator stage is adapted to perform the first stage channel estimation in time domain.

56. Channel estimator according to claim 54 or 55, **characterized in that** the second estimator stage is adapted to perform the second stage channel estimation in frequency domain.

57. Channel estimator according to claim 54 or 55, **characterized in that** the second estimator stage is adapted to perform the second stage channel estimation in time domain.

58. Channel estimator according to claim 45 or 54, **characterized in that** the first stage estimator comprises a first finite impulse response filter adapted to perform the first stage channel estimation in frequency domain is through finite impulse response filtering of received pilot symbols to yield tentative estimates ($\hat{Z}_{\tilde{\ell},\tilde{i}}^{(\mu_1)}$) of received pilot sequences ($\tilde{Z}_{\tilde{\ell},\tilde{i}}^{(\mu_1)}$) in relation to transmission antenna sub-

sets ($\mathcal{A}_{\mu_1}$).

59. Channel estimator according to claim 47 or 56, **characterized in that** the second stage estimator comprises a second finite impulse response filter adapted to perform the second stage channel estimation in frequency domain through finite impulse response filtering of tentative estimates ($\hat{Z}_{\ell,\bar{i}}^{(\mu_1')}$) in relation to antennas in transmission antenna subsets ($\mathcal{A}_{\mu_1}$) to yield estimates of the channel reponse ($\hat{H}_{\ell,i}^{(\mu)}$).

60. Channel estimator according to claim 58 or 59, **characterized in that** the second stage estimator comprises a buffer for executing the second stage channel estimation in frequency domain through finite impulse response filtering with induced time delay for smoothing type filtering.

61. Channel estimator according to one of the claims 58 to 60, **characterized in that** the first finite impulse response filter and/or the second finite impulse response filter is a Wiener type finite impulse response filter.

62. Channel estimator according to claim 46 or 55, **characterized in that** the first stage estimator comprises:

an first inverse fourier transformation unit adapted to execute an inverse fourier transformation of received pilot symbols;

a third finite impulse response filter adapted to execute an optional finite impulse filtering of the inverse fourier transformation result;

a zero padding unit; and

a first fourier transformation unit adapted to execute a fourier transformation of the output of the zero padding unit to yield tentative estimates ($\hat{Z}_{\ell,\tilde{i}}^{(\mu_1)}$) in relation to antennas in transmission antenna subsets ($\mathcal{A}_{\mu_1}$) as first stage channel estimation in time domain.

63. Channel estimator according to claim 48 or 57, **characterized in that** the second stage channel estimator comprises:

a second inverse fourier transformation unit adapted to execute an inverse fourier transformation of tentative estimates ($\hat{Z}_{\ell,\tilde{i}}^{(\mu_1)}$) in relation to antennas in transmission antenna subsets ($\mathcal{A}_{\mu_1}$);

a fourth finite impulse response filter adapted to execute an optional finite impulse filtering of the inverse fourier transformation result;

a zero padding unit;

a second fourier transformation unit adapted to execute a fourier transformation of the output of the zero padding unit to yield estimates of the channel reponse ($\hat{H}_{\ell,i}^{(\mu)}$) as second stage channel estimation in time domain.

64. Channel estimator according to claim 62, **characterized in that** it further comprises an interpolation filter adapted to interpolate the result of finite impulse response filtering before fourier transformation thereof for synthesizing a number of tentative estimates in the time domain corresponding to the number of sub-carriers of the two dimensional data sequence.

65. Channel estimator according to claim 64, **characterized in that** a ratio of the number of tentative estimates in the time domain to the number of received pilot symbols corresponds to pilot spacing in the first dimension of trans-

mission .

66. Channel estimator according to one of the claims 62 to 65, **characterized in that** the second stage estimator is connected to the third finite impulse response filter and adapted to process the result of related finite impulse filtering for determination of the second stage channel estimation.

67. Channel estimator according to one of the claims 35 to 66, **characterized in that** it is operable in a celluar communication system with a frequency reuse factor of one where base stations and related anntenna arrays form the plurality of transmit antennas and such that transmission antenna subsets ($\mathcal{A}_{\mu_1}$) and related transmission antennas are defined in relation to this plurality of transmit antennas.

68. Computer program product directly loadable into the internal memory of a channel estimator for estimating multiple input multiple output transmissio channels in two dimensions comprising software code portions for performing the steps of one of the claims 1 to 34 when the product is run on a processor of the channel estimator for estimating multiple input multiple output transmissio channels in two dimensions.

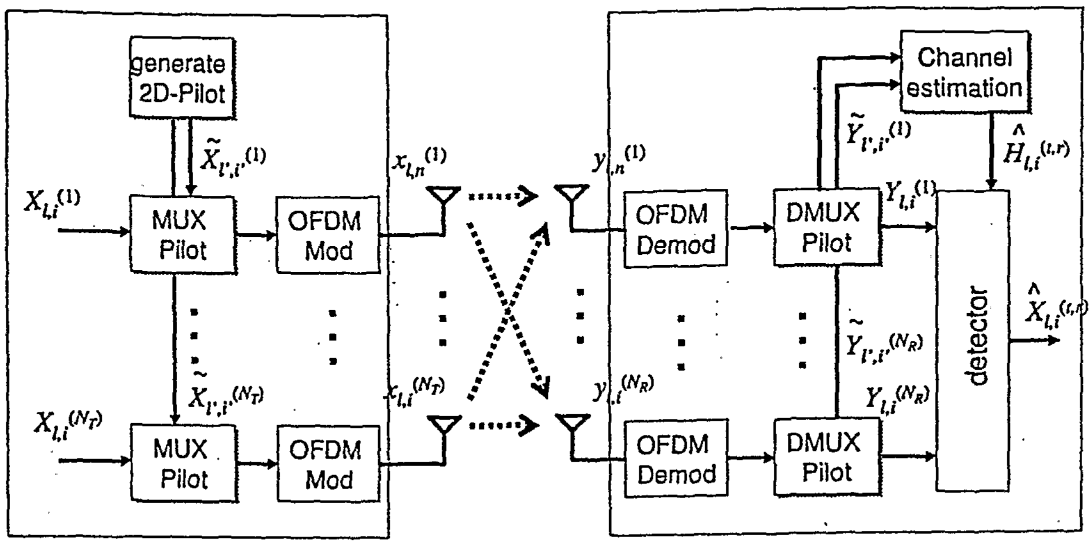

generate 2D-Pilot
MUX Pilot
OFDM Mod
MUX Pilot
OFDM Mod
OFDM Demod
DMUX Pilot
OFDM Demod
DMUX Pilot
Channel estimation
detector

Fig. 1

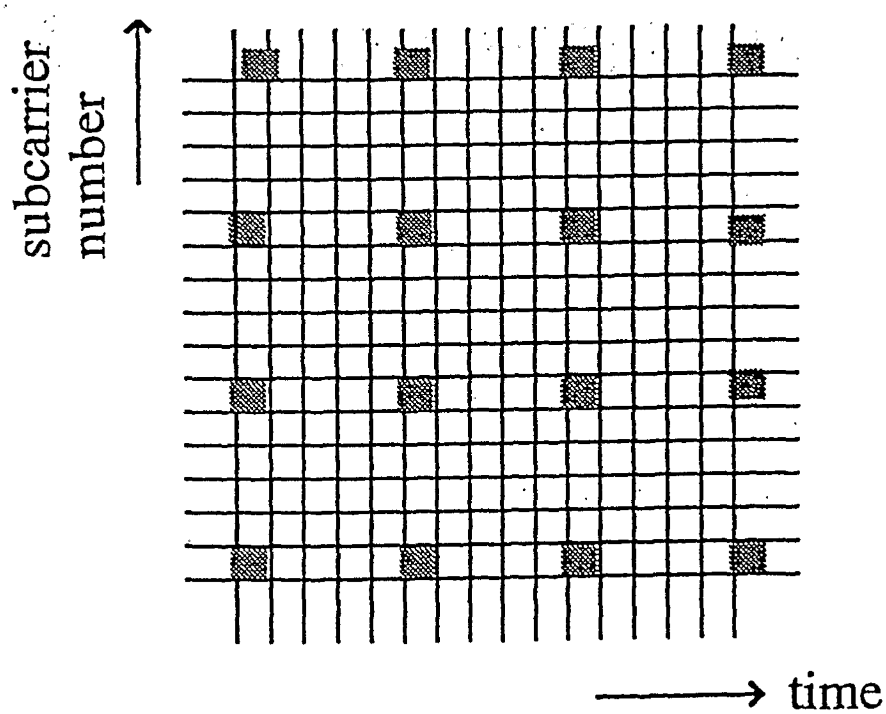

subcarrier number
time

Fig. 2

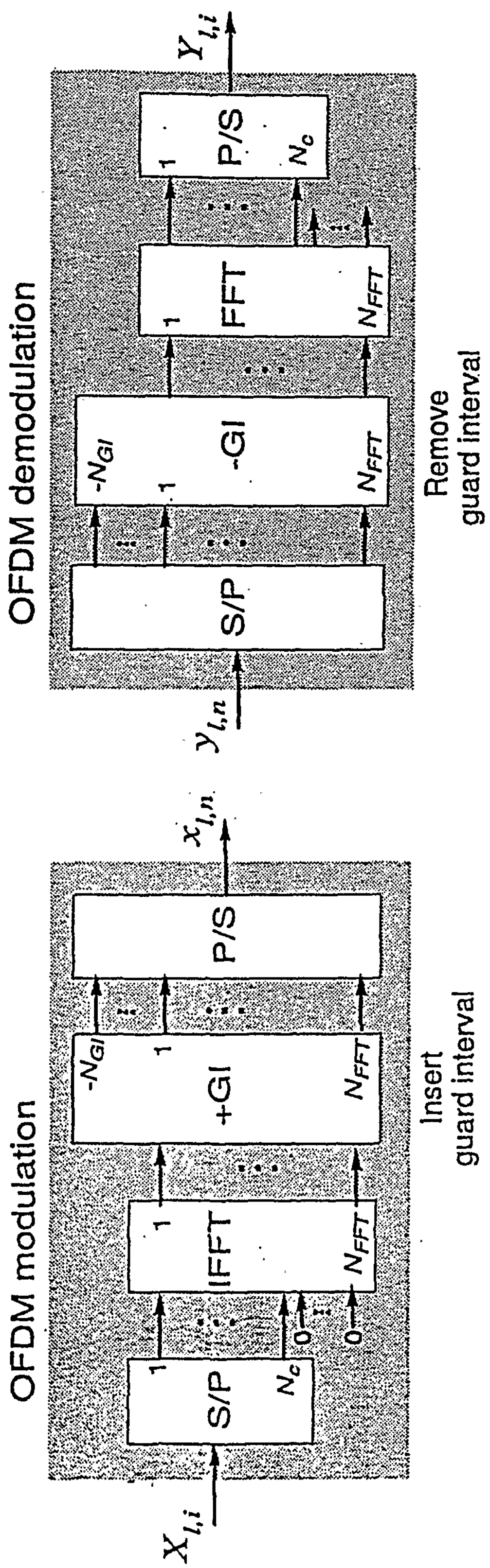

OFDM modulation
$X_{l,i}$
S/P
1
N_c
0
0
IFFT
1
N_{FFT}
+GI
$-N_{GI}$
1
N_{FFT}
Insert
guard interval
P/S
$x_{l,n}$
OFDM demodulation
$y_{l,n}$
S/P
-GI
$-N_{GI}$
1
N_{FFT}
Remove
guard interval
FFT
1
N_{FFT}
P/S
1
N_c
$Y_{l,i}$

Fig 3

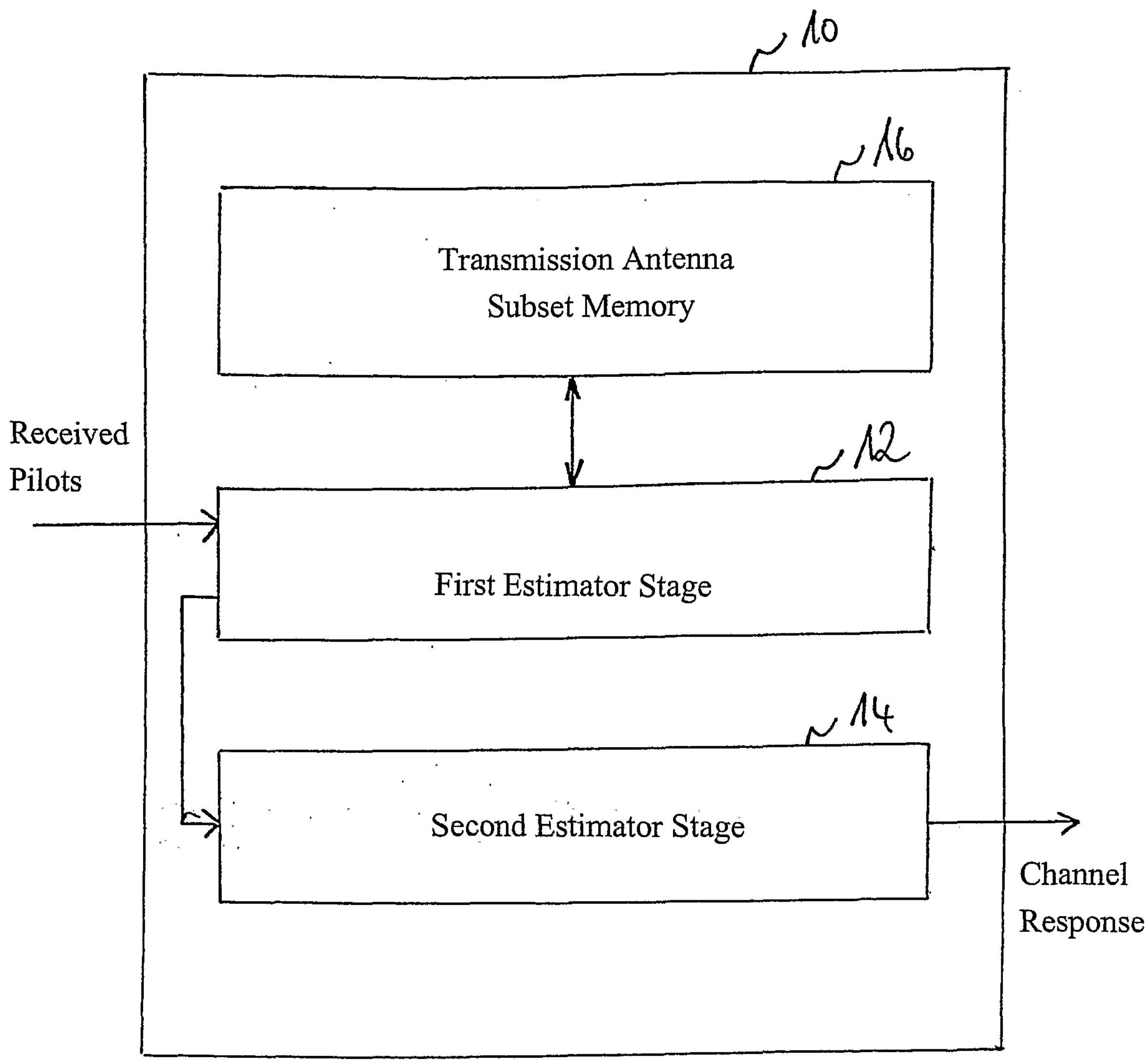

10
16
Transmission Antenna
Subset Memory
Received
Pilots
12
First Estimator Stage
14
Second Estimator Stage
Channel
Response

Fig. 4

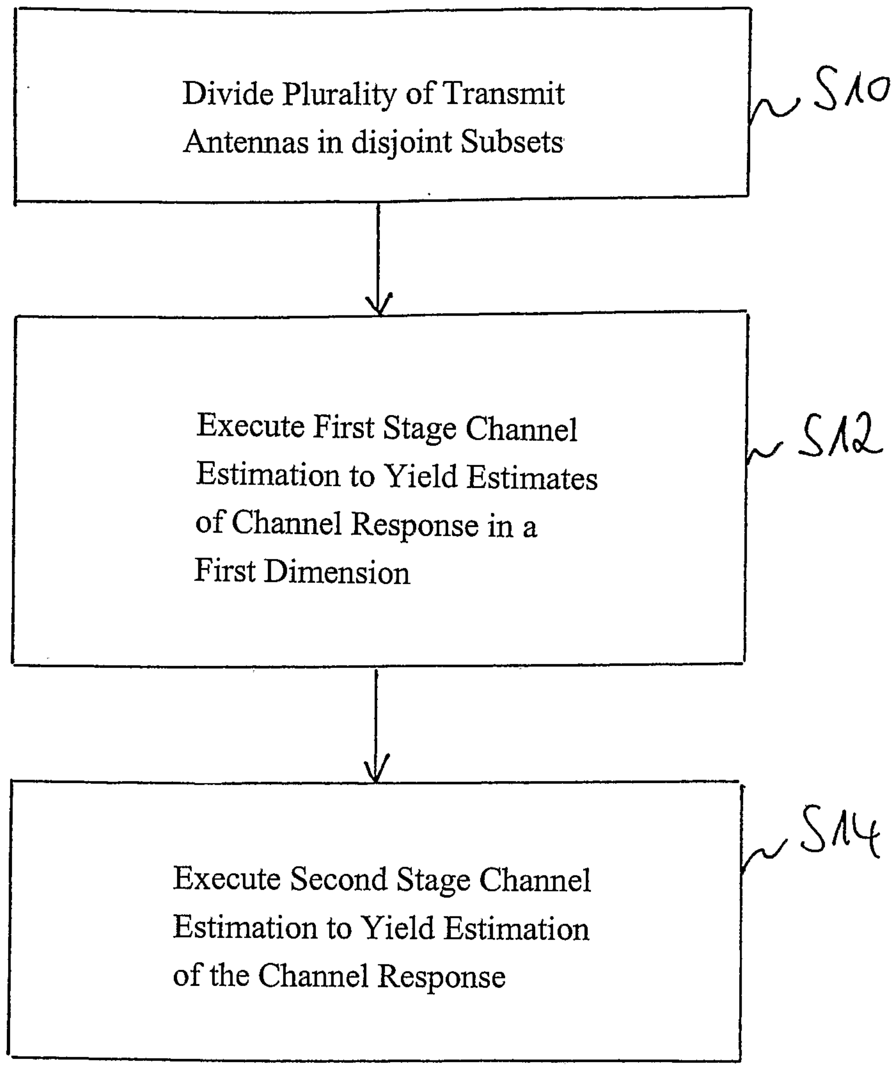

Divide Plurality of Transmit Antennas in disjoint Subsets
S10
Execute First Stage Channel Estimation to Yield Estimates of Channel Response in a First Dimension
S12
Execute Second Stage Channel Estimation to Yield Estimation of the Channel Response
S14

Fig. 5

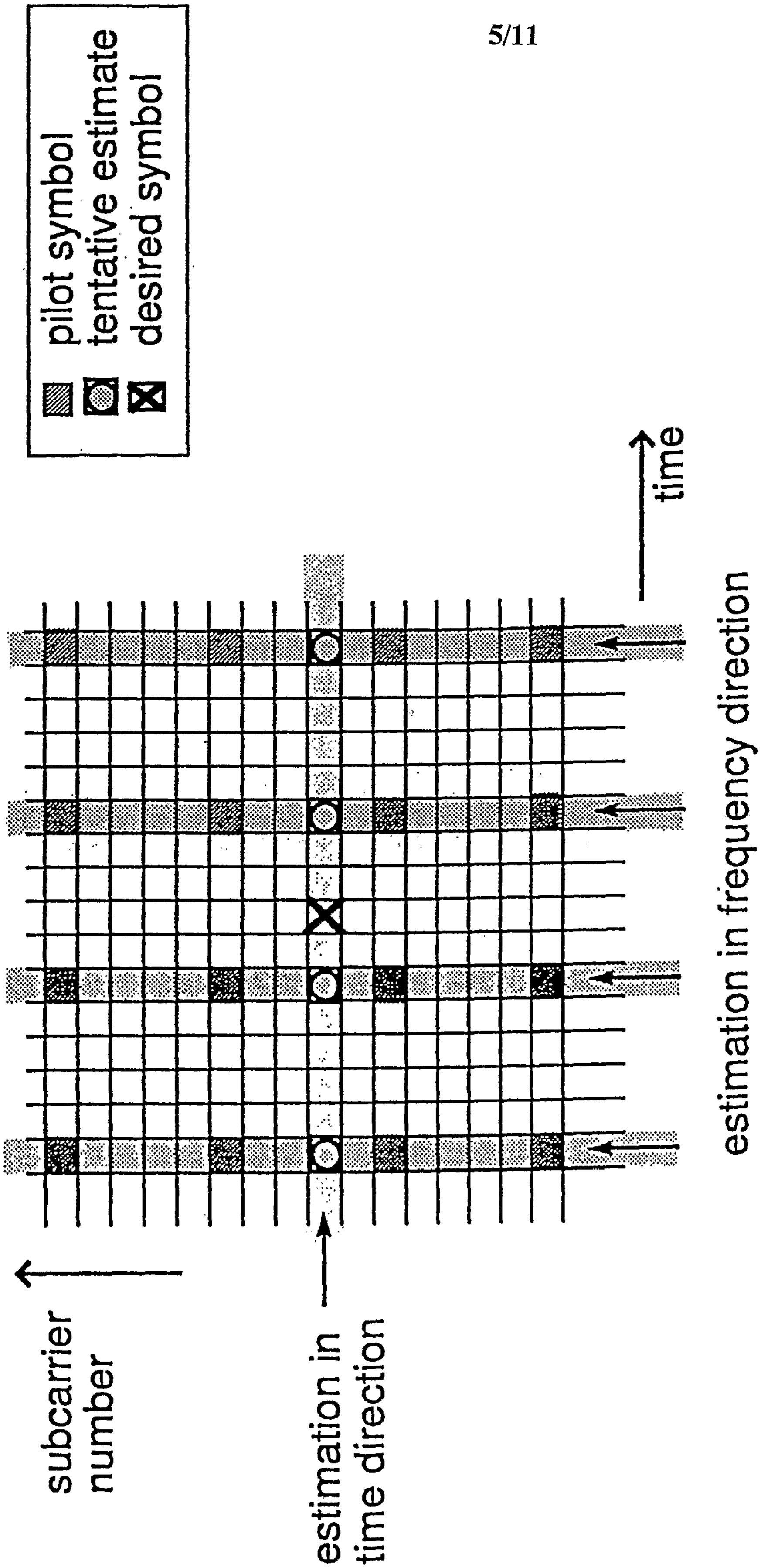

pilot symbol
tentative estimate
desired symbol
time
estimation in frequency direction
subcarrier number
estimation in time direction

Fig. 6

Type I: Channel estimation in frequency direction first

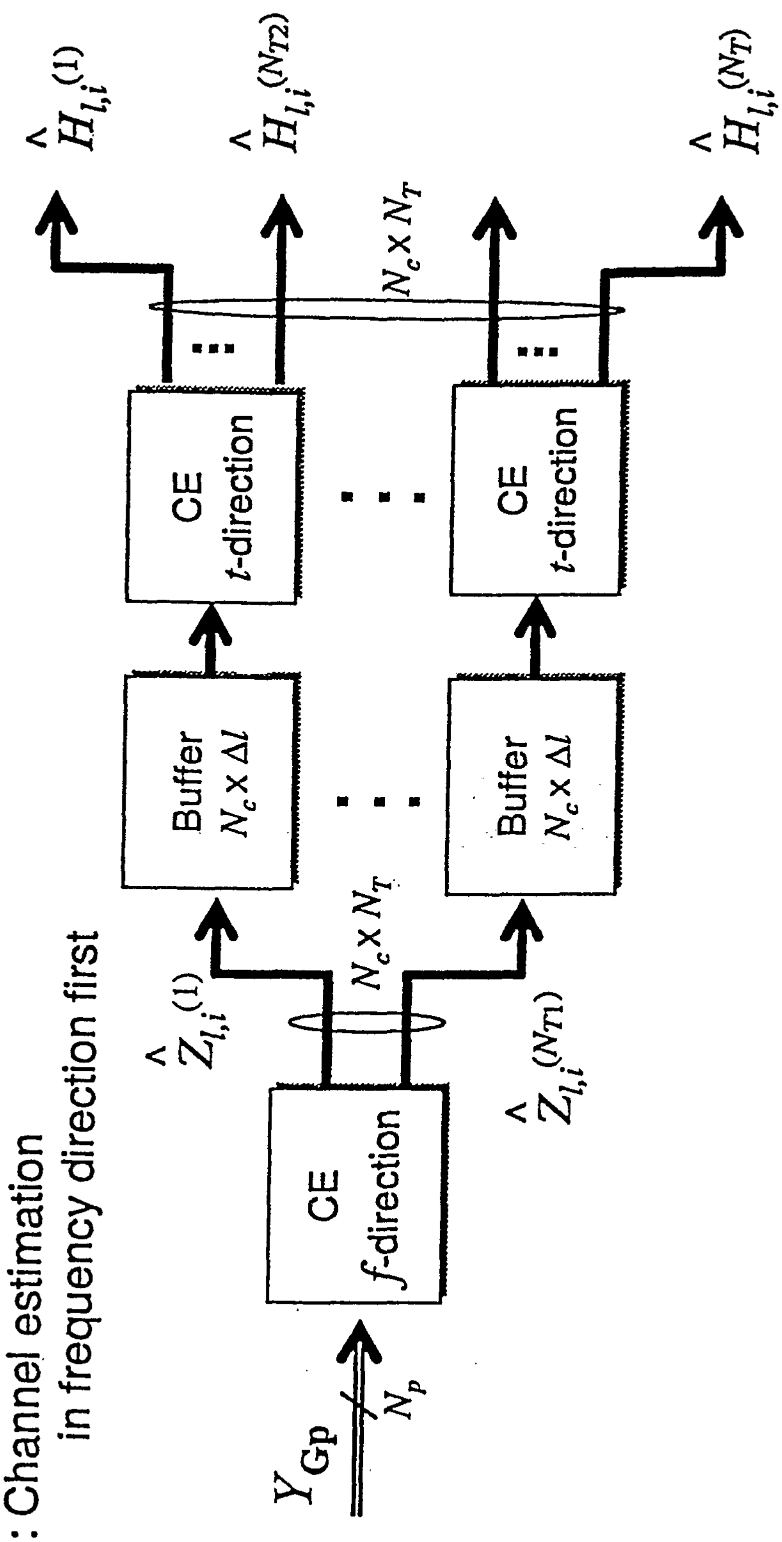


Fig. 7

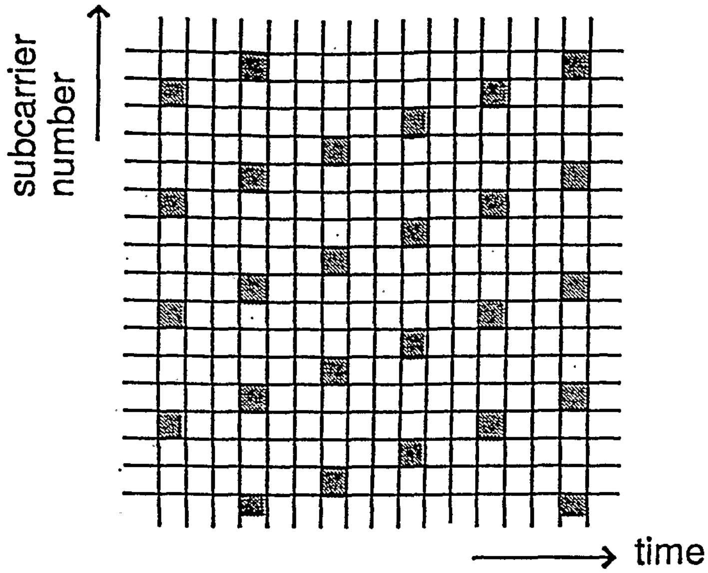


Fig. 8

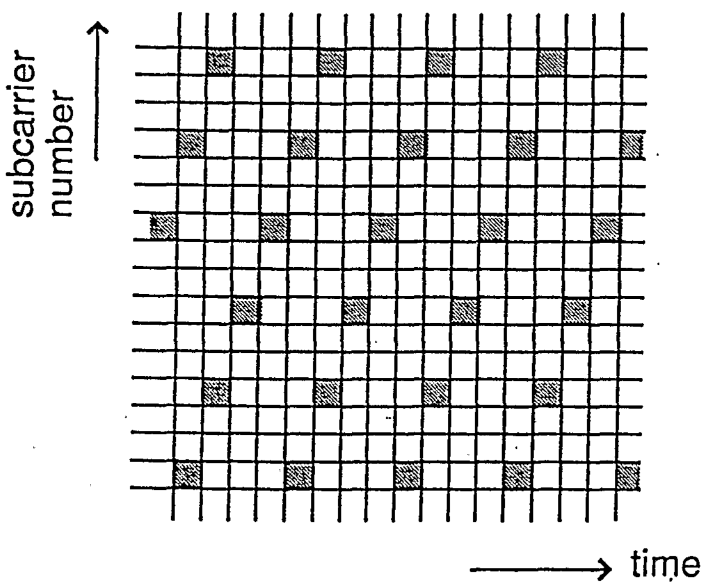


Fig. 9

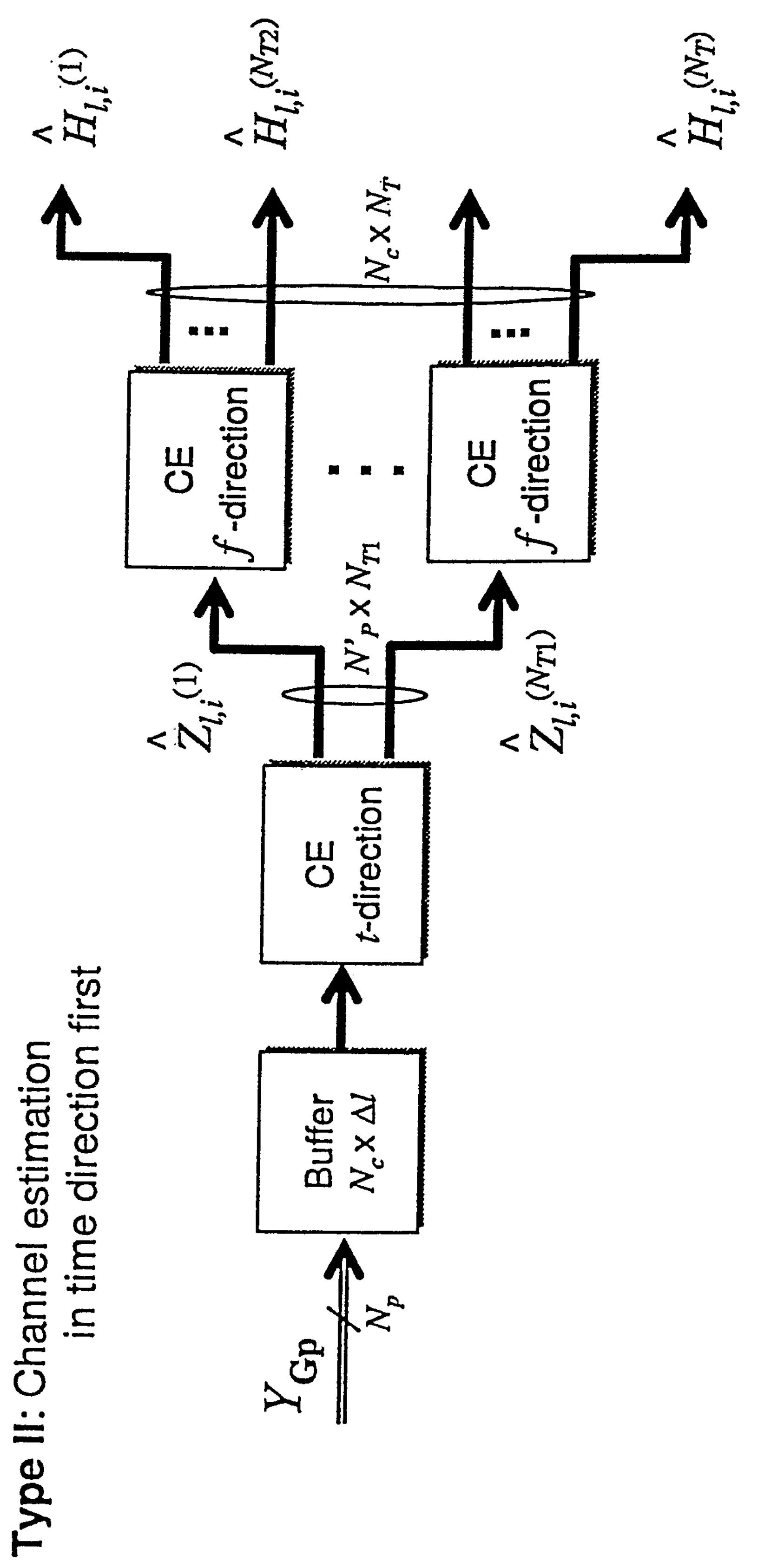

Type II: Channel estimation in time direction first
Y_{Gp}
N_p
Buffer
$N_c \times \Delta l$
CE
t-direction
$\hat{Z}_{l,i}^{(1)}$
$\hat{Z}_{l,i}^{(N_{T1})}$
$N'_p \times N_{T1}$
CE
f-direction
CE
f-direction
$N_c \times N_T$
$\hat{H}_{l,i}^{(1)}$
$\hat{H}_{l,i}^{(N_{T2})}$
$\hat{H}_{l,i}^{(N_T)}$

Fig. 10

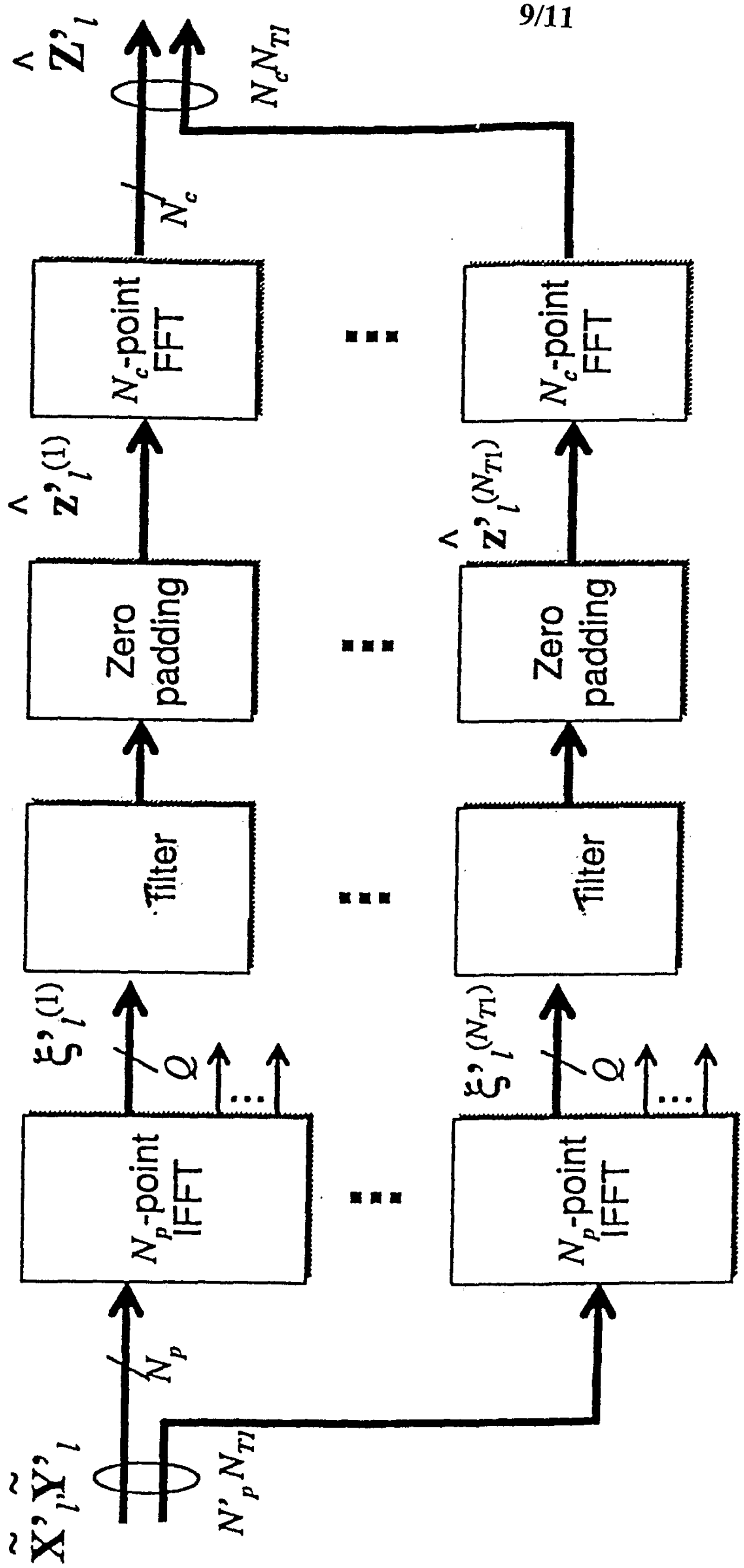


Fig. 11

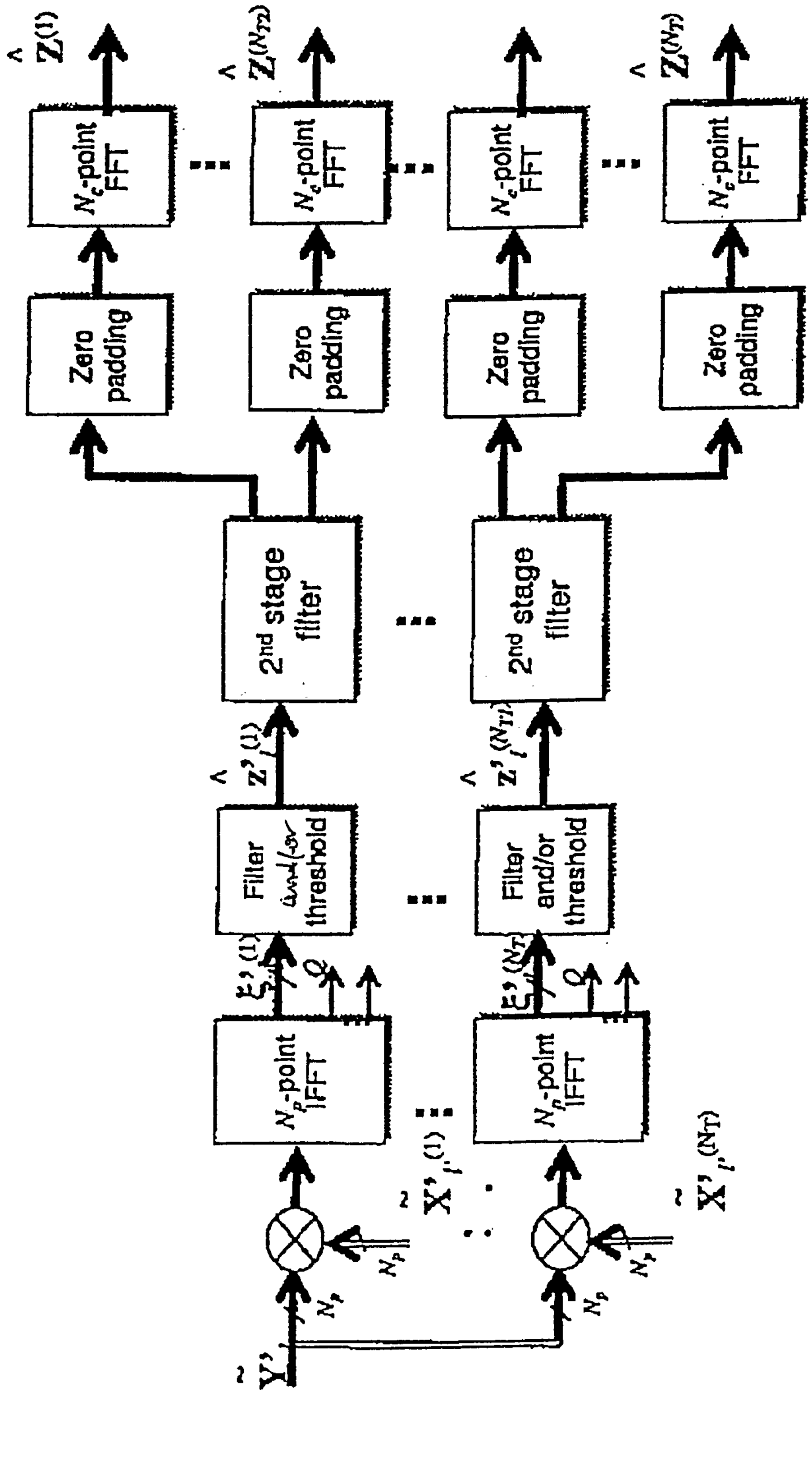

Zero padding
N_c-point FFT
2nd stage filter
Filter and/or threshold
N_p-point IFFT

Fig. 12

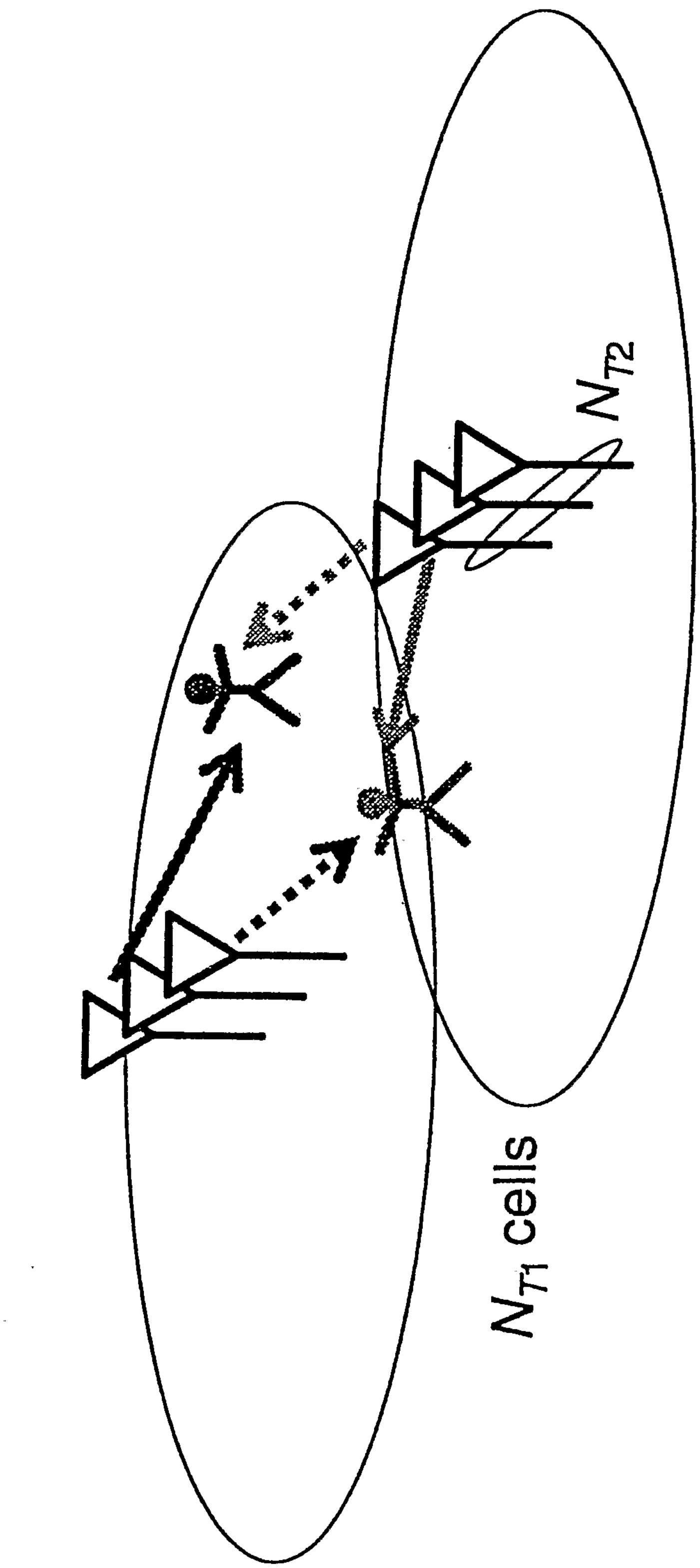


Fig. 13

INTERNATIONAL SEARCH REPORT

International Application No
PCT/EP 03/01495

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L27/26 H04L25/02 H04L1/06

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

EPO-Internal, INSPEC, COMPENDEX, WPI Data, PAJ

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	WON GI JEON ET AL: "Two-dimensional MMSE channel estimation for OFDM systems with transmitter diversity" VTC FALL 2001. IEEE 54TH. VEHICULAR TECHNOLOGY CONFERENCE. PROCEEDINGS. ATLANTIC CITY, NJ, OCT. 7 - 11, 2001, IEEE VEHICULAR TECHNOLGY CONFERENCE, NEW YORK, NY, USA, vol. 1 OF 4. CONF. 54, 7 October 2001 (2001-10-07), pages 1682-1685, XP010562250 ISBN: 0-7803-7005-8 page 1683, right-hand column page 1684, equation (19) section V ——— -/--	1-68

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
27 August 2003	10/09/2003

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31–70) 340–2040, Tx. 31 651 epo nl, Fax: (+31–70) 340–3016	Orozco Roura, C

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
X	EP 1 276 288 A (SIEMENS AG ; SONY GMBH) 15 January 2003 (2003-01-15) page 4, paragraph 15 page 6, paragraph 30 page 16, paragraph 78 ---	1-68
A	HOEHER P ET AL: "TWO-DIMENSIONAL PILOT-SYMBOL-AIDED CHANNEL ESTIMATION" PROCEEDINGS OF THE 1997 IEEE INTERNATIONAL SYMPOSIUM ON INFORMATION THEORY. ISIT '97. ULM, JUNE 29 - JULY 4, 1997, IEEE INTERNATIONAL SYMPOSIUM ON INFORMATION THEORY, NEW YORK, NY: IEEE, US, 29 June 1997 (1997-06-29), page 123 XP000950715 ISBN: 0-7803-3957-6 abstract -----	1-68

INTERNATIONAL SEARCH REPORT

Information on patent family members

International Application No
PCT/EP 03/01495

Patent document cited in search report		Publication date	Patent family member(s)		Publication date
EP 1276288	A	15-01-2003	EP	1276288 A1	15-01-2003